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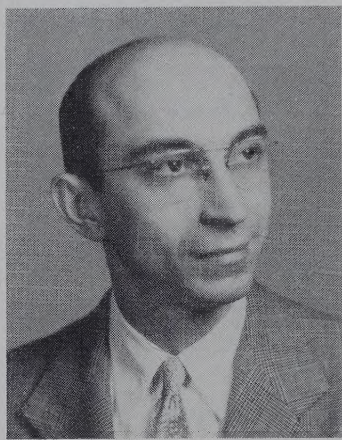

Lotfi A. Zadeh

Lotfi A. Zadeh (S'45—A'47—M'50—SM'56—F'58) was born on February 4, 1921, in Baku, Russia. He attended the American College in Teheran, Iran, and was awarded the B.S. degree in Electrical Engineering by the University of Teheran in 1942. He came to the United States in 1944 and entered Massachusetts Institute of Technology where he received the M.S. degree in 1946.

He joined the staff of Columbia University as an instructor in Electrical Engineering and received the Ph.D. degree from Columbia in 1949. He was promoted to assistant professor in 1950, associate professor in 1953, and became a full professor in 1957. In 1956, he was a member of the Institute for Advanced Study, Princeton, N. J. At present, he is a professor of Electrical Engineering at the University of California in Berkeley.

He has published papers on the subject of time-varying networks, optimal filters, and nonlinear systems, with emphasis on the general principles underlying the transformation of signals and the characterization of input-output relationships.

Dr. Zadeh is Chairman of Subcommissions 6.1, 6.2 of the International Scientific Radio Union, and a member of the American Institute of Electrical Engineers, American Mathematical Society, American Physical Society, Institute of Mathematical Statistics, Association for Computing Machinery, Society for Industrial and Applied Mathematics, Tau Beta Pi, Eta Kappa Nu, and Sigma Xi.

Toward an Institute for Research in Communication Sciences

LOTFI A. ZADEH

Last June I attended in Moscow the 1959 meeting of the A. S. Popov Society, which corresponds roughly to our IRE National Convention. I was very much impressed by the quality and breadth of Soviet work in information theory and related fields, and came back with the conviction that it is imperative to increase the quality and volume of our work in this area by a substantial margin.

What can be done to strengthen our efforts? It seems to me that increasing the level of support for basic research in our universities and industrial laboratories would not by itself yield the desired results. Few of our professors can afford to concentrate on research, most having to sandwich their research activities in between teaching, serving on committees, negotiating contracts, and supervising M.S. and Ph.D. candidates. On the other hand, the atmosphere prevailing in most of our industrial laboratories is generally not conducive to the conduct of basic, nonprofit-oriented, long-term research. Indeed, it is a fact that much of the high quality basic work in our country is turned out by investigators who conduct research on an essentially full-time basis in institute-like organizations within universities and industry.

The key characteristics of such organizations are a) long-term, nonutilitarian attitude toward research, which implies a high degree of freedom for the researcher to work on problems of his own choosing, and insulation from pressures to produce results of immediate practical applicability, b) opportunity for interaction with other researchers, and c) high caliber of staff. A number of well-known institutions both in the United States and abroad have these characteristics, but they are embodied perhaps in their purest form in the Institute for Advanced Study, Princeton, N. J., which since its inception in 1930 has played a very significant role in the development of mathematics in this country.

I should like to suggest that a comparable role in the development of communication theory and related fields could be played by an Institute for Research in Communication Sciences. Such an institute would not necessarily be patterned after the Institute for Advanced Study, nor after the Max Planck Institute in Germany, or the Institute for Automatics and Telemechanics in the Soviet Union. Rather it would be designed to meet the specific needs and interests of workers in the fields of information theory, communication theory, system theory, control theory, circuit theory, automata, biological systems, computation, machine translation of languages, and related fields. It would be concerned with both theoretical and experimental research in these areas. It would not be affiliated with any one educational institution or industrial organization, although it might be advantageous to locate it in the vicinity of a university. It would have extensive library facilities and laboratories. It would be staffed primarily by temporary members, mostly on leave from industrial laboratories and universities. And it would be supported by the Federal government or a foundation, or a group of universities and industrial laboratories. A case in point, though not necessarily a model to follow closely, is the Brookhaven National Laboratory, Upton, Long Island.

Such an institute would supplement, rather than supplant, the current research effort in universities and industry. It would provide a place where professors and industrial researchers could spend a year or two working full-time on their problems, interacting with one another in an atmosphere free from the usual pressures and harrassments. Perhaps this sounds like a dream, but it is a dream worth striving for.

What are the chances of getting the necessary financial support for such an institute? Clearly, it would take a concerted effort of many people to implement the idea, and the prospects for success at this time are rather dim. Nonetheless, whoever feels strongly that the creation of the Institute for Research in Communication Sciences is in the national interest has the responsibility to press for its establishment regardless of how difficult the task might be. History is replete with instances where eventual success has been achieved against even greater odds.

Optimum System Theory Using a General Bayes Criterion*

V. S. PUGACHEV†

Summary—**An extension of the general method of obtaining an optimum system developed by the author[1] is given to include the case of nonlinear dependence of the observed function on signal parameters. The method affords effective determining of optimum systems designed for the detection and reproduction of signals in the presence of noise using various practically adequate criteria.**

I. Formulation of the Problem

MOST of the problems of obtaining an algorithm of an optimum system designed for detecting and reproducing signals in the presence of interferences may be formulated as follows.

For a random function

$$Z(t) = \varphi(t, U) + X(t) \tag{1}$$

being observed in a certain domain T of the argument t, it is necessary to find an estimate $W^*(s)$ of a signal $W(s)$ of the form

$$W(s) = \psi(s, U, Y(s)). \tag{2}$$

In (1) and (2) φ and ψ are definite functions, U is a random vector, and $X(t)$ and $Y(s)$ are random functions. Arguments t and s in problems of information theory and automatic control theory are the current time in the observation interval T, and the end of this interval, respectively. The domain T of the variation of t, in which the function Z is observed, may be a discrete set of points, an interval $s - T \leq t \leq s$, or a set of nonoverlapping intervals. In other problems to which the theory developed below is applicable, arguments t and s may be arbitrary scalar or vector variables varying over arbitrary sets T and S, respectively.

The theory also includes the case in which the observed function $Z(t)$ and the signal $W(s)$ are vector random functions. To apply the theory to this case it is sufficient to consider the components $Z_\nu(t)$, $W_\mu(s)$ as scalar random functions of the arguments t, s and the integers ν, μ. For this reason the theory developed below is applicable as well to many other problems such as, for example, weather prediction, calculating the elements of the motion of satellites and cosmic rockets using observed data, etc.

Different criteria of optimality are adequate in differe practical problems. Yet most practically adequate criter belong to the class of Bayes criteria of the form

$$M[l(W, W^*)] = \min, \tag{3}$$

where a loss function $l(W, W^*)$ may be an arbitrar function or functional of the signal W and its estima W^*, and the symbol $M[X]$ denotes the mathematic expectation of the variable X.[2-4] The loss function may also depend upon certain indefinite parameters chose to satisfy the given complementary conditions.

We shall obtain here the solution of the problem state above for any functions φ, ψ and l under the followi general conditions:

1) U is a finite-dimensional random vector with known probability density $f(u)$.
2) $[X(t), Y(s)]$ is a normally distributed random vect function independent of the random vector U.
3) For each realization z of the observed random fun tion Z there exists such a function w^* that for a function w the inequality

$$M[l(W, w^*) \mid z] \leq M[l(W, w) \mid z] \tag{4}$$

   is valid. $M[X \mid z]$ denotes here the condition expectation of X for a given realization z of This condition is sufficient for the existence of su a nonrandom operator A that the estimate

$$W^* = AZ \tag{5}$$

   provides the minimum average risk (3). This cond tion implies certain limitations upon the class admissible functions φ, ψ and l and is satisfied almost all practical problems.
4) For all values of u the function $\varphi(t, u)$ is represen able by the series of eigenfunctions of the rand function $X(t)$ in the domain T.
5) The realization of the random function Y is co pletely determined if the corresponding realizati of the random function X is known.

* Manuscript received by the PGIT, July 6, 1959.

† Inst. of Automatics and Telemechanics, U.S.S.R. Academy of Sciences, Moscow.

[1] V. S. Pugachev, "Determination of an optimum system using a general criterion," *Avtomat. i Telemekh.*, vol. 19, pp. 519–539; June, 1958 (in Russian); also *Automation and Remote Control*, same issue (in English).

[2] A. Wald, "Statistical Decision Functions," John Wiley a Sons, Inc., New York, N. Y.; 1950.

[3] D. Van Meter, and D. Middleton, "Modern statistical proaches to reception in communication theory," IRE TRA on Information Theory, vol. IT-4, pp. 119–145; September, 19

[4] D. Middleton, and D. Van Meter, "Detection and extract of signals in noise from the point of view of statistical decis theory, Parts I and II," SIAM J., vol. 3, pp. 192–253, Decemb 1955 and vol. 4, pp. 86–119; June, 1956.

II. The Solution of the Problem

To solve the problem we represent the vector random function $[X(t), Y(s)]$ by some canonical expansion[5-7]

$$X(t) = \sum_\nu V_\nu x_\nu(t), \qquad Y(s) = \sum_\nu V_\nu y_\nu(s), \tag{6}$$

where V_ν are uncorrelated random variables and $x_\nu(t)$, $y_\nu(s)$ are nonrandom coordinate functions. The expansions (6) may always be found in an infinite variety of ways for any finite domains T, $S(t \varepsilon T,\ s \varepsilon S)$. In particular, real random functions X, Y are always representable by expansions (6) with real coordinate functions x_ν, y_ν and coefficients V_ν. Once real expansions (6) are obtained, it is possible to find a system of real linear functionals $\Omega^{(\nu)}$, acting upon the values of functions of the variable t varying in the domain T, satisfying

$$\Omega^{(\nu)} x_\mu = \delta_{\nu\mu}, \tag{7}$$

$$x_\nu(t) = \frac{1}{D_\nu} \Omega_\tau^{(\nu)} K_x(t, \tau), \tag{8}$$

where D_ν is the variance of the random variable V_ν, K_x is the correlation function (covariance) of the random function X and $\delta_{\nu\mu}$ is Kronecker's delta. Then the random coefficients V_ν contained in the expansion (6) of X are representable by[5,6]

$$V_\nu = \Omega^{(\nu)} X. \tag{9}$$

In the case of a discrete observation domain T, the functionals $\Omega^{(\nu)}$ are linear combinations of values of corresponding functions in points of the domain T:

$$\Omega^{(\nu)} \gamma = \sum_{t_k \varepsilon T} a_{\nu k} \gamma(t_k). \tag{10}$$

In the case of a continuous observation domain T and scalar function Z, the functionals $\Omega^{(\nu)}$ are integrals extending over the domain T:

$$\Omega^{(\nu)} \gamma = \int_T a_\nu(t) \gamma(t)\, dt. \tag{11}$$

Finally in the case of a continuous observation domain T and vector function Z, the functionals $\Omega^{(\nu)}$ are sums of integrals:

$$\Omega^{(\nu)} \gamma = \sum_p \int_T a_{\nu p}(t) \gamma_p(t)\, dt. \tag{12}$$

[5] V. S. Pugachev, "Use of canonical expansions of random functions to derive optimal linear systems," *Avtomat. i Telemekh.*, vol. 17, pp. 489–499; June, 1956 (in Russian); also *Automation and Remote Control*, same issue (in English).

[6] V. S. Pugachev, "Theory of Random Functions and its Application to Automatic Control Problems," State Publishing House of Theoretical-Technical Literature, Moscow; 1957. (In Russian.)

[7] Expectations of random functions X, Y may clearly be assumed without loss of generality to be equal identically to zero.

Assumption 4) of the preceding section means that for all values of the vector u the function $\varphi(t, u)$ may be represented in the domain T by the expansion

$$\varphi(t, u) = \sum_\nu D_\nu \alpha_\nu(u) x_\nu(t) \qquad (t \varepsilon T). \tag{13}$$

where

$$D_\nu \alpha_\nu(u) = \Omega^{(\nu)} \varphi(t, u). \tag{14}$$

Assumption 5) of Section I means that the expansion (6) of the random function Y contains only such random variables V_ν as are also present in the expansion (6) of the random function X. That means also that all random variables V_ν are representable by (9).

Using the fact that V_ν are independent normally distributed random variables, being linear transforms of the normally distributed random function X, we obtain the following expression for the conditional expectation of the loss function l with respect to the observed function Z:

$$M[l(W, W^*) \mid Z] = c \int_{-\infty}^{\infty} l(\psi(s, u, A^{(0)} Z - \omega(u)), W^*) \cdot f(u) \exp\{L(u)Z - \tfrac{1}{2}\beta(u)\}\, du, \tag{15}$$

where $A^{(0)}$ and $L(u)$ are linear operators defined by

$$A^{(0)} = \sum_\nu y_\nu \Omega^{(\nu)}, \qquad L(u) = \sum_\nu \alpha_\nu(u) \Omega^{(\nu)}, \tag{16}$$

$\omega(u)$ and $\beta(u)$ are functions defined by

$$\left.\begin{aligned} \omega(u) &= \sum_\nu D_\nu \alpha_\nu(u) y_\nu(s) = A_t^{(0)} \varphi(t, u), \\ \beta(u) &= \sum_\nu D_\nu \alpha_\nu^2(u) = L(u)_t \varphi(t, u) \end{aligned}\right\}, \tag{17}$$

and c is a certain constant depending on the observed function Z and independent of W^*.

Thus the problem of finding an optimal estimate W^* of the signal W is reduced to minimization of the integral (15) after calculating linear operators $A^{(0)}$ and $L(u)$ and functions $\omega(u)$ and $\beta(u)$.

It is easily verified by direct substitution that linear operators $A^{(0)}$ and $L(u)$ satisfy

$$\begin{aligned} A_\tau^{(0)} K_x(t, \tau) &= K_{yx}(s, t), \\ L(u)_\tau K_x(t, \tau) &= \varphi(t, u) \qquad (t \varepsilon T), \end{aligned} \tag{18}$$

where, in addition to previous notations, $K_{yx}(s, t)$ is the cross-correlation function of random functions Y and X. Thus series (16) determine solutions of (18).[5,6] If (18) may be solved in finite form by some procedure, then operators $A^{(0)}$, $L(u)$ and functions $\omega(u)$, $\beta(u)$ may be found without using infinite series.

In the case of a discrete observation domain T, (18) are systems of linear algebraic equations. In the case of a continuous observation domain T, (18) are Fredholm

integral equations of the first kind if Z is a scalar function, and systems of Fredholm integral equations of the first kind if Z is a vector function. To solve (18) in finite form the method of integral canonical representations of random functions may be used.[8,9]

The sufficient conditions for convergence in mean square sense of the series

$$\sum_{\nu} y_{\nu}\Omega^{(\nu)}Z \quad \text{and} \quad \sum_{\nu} \alpha_{\nu}(u)\Omega^{(\nu)}Z \tag{19}$$

are the existence of finite variance of the random function Y and the convergence for all u of the second series (17).[5,6]

Minimum conditional expectations (15) corresponding to successive partial sums of series (16) will converge in probability to the minimum value of the expectation $M[l(W, W^*) \mid Z]$ in this case for a wide class of functions l, φ and ψ.

The method of obtaining the optimum system algorithm, developed above, is also applicable to the case when some components (or all) of the vector U are nonrandom unknown quantities which may assume any values. In this case $f(u)$ represents the joint probability density of random components of the vector U (unity if all components of the vector U are nonrandom).

III. Some Applications of the Method

Defining the loss function l by

$$l(w, w^*) = \begin{cases} 0 \text{ if } w = 0, w^* \le c \text{ or } w \ne 0, w^* > c, \\ 1 \text{ if } w \ne 0, w^* \le c, \\ \lambda \text{ if } w = 0, w^* > c, \end{cases} \tag{20}$$

the method developed in the preceding section yields the general solution of the problem of detecting a signal in noise. The constant λ is equal to unity if the ideal observer criterion is used, and is determined by a given value of the probability of a false alarm if the Neyman-Pearson criterion is used. Yet the loss function is reduced to a form (20) in the case of arbitrary chosen weights of correct decisions and errors of the two types.

If φ is a linear function of components $U_1, \cdots, U_N$ of the vector U,

$$\varphi(t, U) = \sum_{k=1}^{N} U_k\varphi_k(t), \tag{21}$$

the method yields an explicit expression of the optimum system operator.[1]

The function ψ being also linear in $U_1, \cdots, U_N$, Y random components of the vector U being normally distributed and the loss function depending only on the difference $W^* - W$, the method yields a linear optimum system and the form of the loss function affects only a systematic part of the optimum system output, independent of the specific realization of the input Z. If, in addition the loss function l is a nondecreasing function of the modulus of the difference $W^* - W$, then the optimum system is completely independent of the specific choice of the loss function l of that form and coincides with that determined by the least mean square criterion for this case.[1] These statements are also valid in the case when the observation domain T is the interval $s_0 - T \le t \le s_0$ and the loss function l is a functional of the form

$$l(W, W^*) = \int_S \sigma(s, \Delta, \Delta', \cdots, \Delta^{(l)})\, ds, \tag{22}$$

where

$$\Delta = \Delta(s) = W^*(s) - W(s), \tag{23}$$

σ is any function of the arguments indicated, and S is any domain of the variation of the argument s.

In the case where $Z(t)$ is a scalar function, U is a nonrandom vector, the observation domain T is the interval $s - T \le t \le s$, $W = U$, and the loss function is determined by

$$l(w, w^*) = \text{const} - \delta(w^* - w) \tag{24}$$

where δ is the Dirac delta function, the general techniques derived in the preceding section yield the equations determining the estimates of signal parameters obtained by Slepian by application of the usual maximum likelihood method.[10]

The method may also be applied to find an estimate of an infinite-dimensional normally distributed vector U using the maximum likelihood criterion. There results in this case the infinite system of equations determining estimates of components of the vector U, similar to Slepian's system of equations. Using this fact, it is possible to obtain an algorithm for obtaining the maximum likelihood estimate of a normally distributed random function $W(t)$, upon which the observed function Z depends

$$Z(t) = \mu(t, W(t)) + X(t). \tag{25}$$

For this purpose it is sufficient to express the random function $W(t)$ by some canonical expansion:

$$W(t) = \sum_{k} U_k w_k(t). \tag{26}$$

Then, departing from the equations, determining estimates of independent random variables U_k, the system of equa

[8] V. S. Pugachev, "Integral canonic representation of random functions and their application in deriving optimal linear systems," *Avtomat. i Telemekh.*, vol. 18, pp. 971–984; November, 1957 (in Russian), also *Automation and Remote Control*, same issue (in English).

[9] V. S. Pugachev, "Method of solving the basic integral equation of the statistical theory of optimum systems in closed form," *Prikladnaia Matemat. i Mekh.*, vol. 23, pp. 3–14; January, 1959.

[10] D. Slepian, "Estimation of signal parameters in the presence of noise," IRE Trans. on Information Theory, vol. IT-3, pp. 68–89; March, 1954.

ons for the maximum likelihood estimate of the random ınction, $W(t)$, which was previously obtained by Youla ›llows.[11] There also results another system of equations etermining that estimate of the random function $W(t)$ hich is equivalent to Youla's system of equations.

The general method discussed in the preceding section lso yields the solution of many other problems of optimum ystem theory and similar problems of many other ranches of science.

IV. Concluding Remarks

We see that the general method presented above yields he effective solution of various problems of applied staistical decision theory, particularly, of the statistical heory of signal detection and reproduction.

The algorithm given by the method for obtaining a signal estimate may be used as a base for real system design. The main difficulty which arises in the practical realization of such systems is generally the absence of necessary data characterizing the *a priori* distribution of the signal parameter U [*i.e.*, the probability density $f(u)$]. To avoid this difficulty, the same algorithm may be applied to obtain the signal parameter U estimate in each cycle of the system acting, and to construct for each cycle an estimate for $f(u)$ using estimates of U obtained in all previous cycles. Using such estimates of $f(u)$ for each cycle of the system acting, we obtain a "self-learning" system which will be nearer and nearer with each new cycle to the true optimum system corresponding to the true probability distribution of the vector U.[12]

[11] D. Youla, "The use of the method of maximum likelihood in stimating continuous-modulated intelligence which has been orrupted by noise," IRE Trans. on Information Theory, ol. IT-3, pp. 90–105; March, 1954.

[12] K. Winkelbauer, "Experience in Games of Strategy and In Statistical Decision," *Trans. of the First Prague Conf. on Information Theory, Statistical Decision Functions and Random Processes*, Liblice, November 28–30, 1956, Czech. Acad. of Sc., Prague, pp. 297–354, 1957.

Quantizing for Minimum Distortion*

JOEL MAX†

***Summary*—This paper discusses the problem of the minimization f the distortion of a signal by a quantizer when the number of utput levels of the quantizer is fixed. The distortion is defined as he expected value of some function of the error between the input nd the output of the quantizer. Equations are derived for the paramters of a quantizer with minimum distortion. The equations are ot soluble without recourse to numerical methods, so an algorithm s developed to simplify their numerical solution. The case of an ıput signal with normally distributed amplitude and an expected quared error distortion measure is explicitly computed and values f the optimum quantizer parameters are tabulated. The optimization f a quantizer subject to the restriction that both input and output evels be equally spaced is also treated, and appropriate parameters re tabulated for the same case as above.**

* Manuscript received by the PGIT, September 25, 1959. This work was performed by the Lincoln Lab., Mass. Inst. Tech., Lexington, Mass., with the joint support of the U. S. Army, Navy, nd Air Force.

† Lincoln Lab., Mass. Inst. Tech., Lexington, Mass.

IN MANY data-transmission systems, analog input signals are first converted to digital form at the transmitter, transmitted in digital form, and finally reconstituted at the receiver as analog signals. The resulting output normally resembles the input signal but is not precisely the same since the quantizer at the transmitter produces the same digits for all input amplitudes which lie in each of a finite number of amplitude ranges. The receiver must assign to each combination of digits a single value which will be the amplitude of the reconstituted signal for an original input anywhere within the quantized range. The difference between input and output signals, assuming errorless transmission of the digits, is the quantization error. Since the digital transmission rate of any system is finite, one has to use a quantizer which sorts the input into a finite number of ranges, N. For a given N, the system is described by specifying the end

points, x_k, of the N input ranges, and an output level, y_k, corresponding to each input range. If the amplitude probability density of the signal which is the quantizer input is given, then the quantizer output is a quantity whose amplitude probability density may easily be determined as a function of the x_k's and y_k's. Often it is appropriate to define a distortion measure for the quantization process, which will be some statistic of the quantization error. Then one would like to choose the N y_k's and the associated x_k's so as to minimize the distortion. If we define the distortion, D, as the expected value of $f(\epsilon)$, where f is some function (differentiable), and ϵ is the quantization error, and call the input amplitude probability density $p(x)$, then

$$D = E[f(s_{\text{in}} - s_{\text{out}})]$$
$$= \sum_{i=1}^{N} \int_{x_i}^{x_{i+1}} f(x - y_i)p(x)\,dx$$

where $x_{N+1} = \infty$, $x_1 = -\infty$, and the convention is that an input between x_i and x_{i+1} has a corresponding output y_i.

If we wish to minimize D for fixed N, we get necessary conditions by differentiating D with respect to the x_i's and y_i's and setting derivatives equal to zero:

$$\frac{\partial D}{\partial x_j} = f(x_j - y_{j-1})p(x_j) - f(x_j - y_j)p(x_j) = 0 \qquad j = 2, \cdots, N \qquad (1)$$

$$\frac{\partial D}{\partial y_j} = -\int_{x_j}^{x_{j+1}} f'(x - y_j)p(x)\,dx = 0 \qquad j = 1, \cdots, N \qquad (2)$$

(1) becomes (for $p(x_j) \neq 0$)

$$f(x_j - y_{j-1}) = f(x_j - y_j) \qquad j = 2, \cdots, N \qquad (3)$$

(2) becomes

$$\int_{x_j}^{x_{j+1}} f'(x - y_j)p(x)\,dx = 0 \qquad j = 1, \cdots, N. \qquad (4)$$

We may ask when these are sufficient conditions. The best answer one can manage in a general case is that if all the second partial derivatives of D with respect to the x_i's and y_i's exist, then the critical point determined by conditions (3) and (4) is a minimum if the matrix whose ith row and jth column element is

$$\left.\frac{\partial^2 D}{\partial p_i\, \partial p_j}\right|_{\text{critical point}},$$

where the p's are the x's and y's, is positive definite. In a specific case, one may determine whether or not the matrix is positive definite or one may simply find all the critical points (*i.e.*, those satisfying necessary conditions) and evaluate D at each. The absolute minimum must be at one of the critical points since "end points" can be easily ruled out.

The sort of f one would want to use would be a goo metric function, *i.e.*, $f(x)$ is monotonically nondecreasin

$$f(0) = 0$$
$$f(x) = f(-x).$$

If we require that $f(x)$ be *monotonically increasing* (with x then (1) implies

$$|x_j - y_{j-1}| = |x_j - y_j| \qquad j = 2, \cdots, N$$

which implies (since y_{j-1} and y_j should not coincide) tha

$$x_j = (y_j + y_{j-1})/2 \qquad j = 2, \cdots, N$$

(x_j is halfway between y_j and y_{j-1}).

We now take a specific example of $f(x)$ to furthe illuminate the situation.

Let $f(x) = x^2$

(3) implies

$$x_j = (y_j + y_{j-1})/2 \quad \text{or} \quad y_j = 2x_j - y_{j-1} \qquad j = 2, \cdots, N, \qquad (5$$

(4) implies

$$\int_{x_j}^{x_{j+1}} (x - y_j)p(x)\,dx = 0 \qquad j = 1, \cdots, N. \qquad (6$$

That is, y_j is the centroid of the area of $p(x)$ between x and x_{j+1}.

Because of the complicated functional relationship which are likely to be induced by $p(x)$ in (6), this i not a set of simultaneous equations we can hope to solv with any ease. Note, however, that if we choose y_1 cor rectly we can generate the succeeding x_j's and y_j's b (5) and (6), the latter being an implicit equation for x_{j+} in terms of x_j and y_j.

A method of solving (5) and (6) is to pick y_1, calculat the succeeding x_j's and y_j's by (5) and (6) and then i y_N is the centroid of the area of $p(x)$ between x_N and ∞ y_1 was chosen correctly. (Of course, a different choice i appropriate to each value of N.) If y_N is not the appro priate centroid, then of course y_1 must be chosen agair This search may be systematized so that it can be per formed on a computer in quite a short time.[1]

This procedure has been carried out numerically on th IBM 709 for the distribution $p(x) = 1/\sqrt{2\pi}\, e^{-x^2/2}$, unde the restriction that $x_{N/2+1} = 0$ for N even, and $y_{(N+1)/2} =$ for N odd. This procedure gives symmetric results, *i.e*

[1] Obtaining *explicit* solutions to the quantizer problem for nontrivial $p(x)$ is easily the most difficult part of the problem The problem may be solved analytically where $p(x) = 1/\sqrt{2\pi}\, e^{-x^2}$ only for $N = 1$, $N = 2$. For $N = 1$, $x_1 = -\infty$, $y_1 = 0$, $x_2 = +\infty$ For $N = 2$, $x_1 = -\infty$, $y_1 = -\sqrt{2/\pi}$, $x_2 = 0$, $y_2 = \sqrt{2/\pi}$, $x_3 = +\infty$ ($\sqrt{2/\pi}$ is the centroid of the portion of $1/\sqrt{2\pi}\, e^{-x^2/2}$ between th origin and $+\infty$.) For $N \geq 3$, some sort of numerical estimation required. A somewhat different approach, which yields resul somewhat short of the optimum, is to be found in V. A. Gamas "Quantization of signals with non-uniform steps," *Electrosvyaz*, vo 10, pp. 11–13; October, 1957.

a signal amplitude x is quantized as y_k, then $-x$ is ıantized as $-y_k$. The answers appear in Table I on ıge 11.

An attempt has been made to determine the functional ›pendence of the distortion on the number of output vels. A log-log plot of the distortion vs the number of ıtput levels is in Fig. 1. The curve is not a straight line. he tangent to the curve at $N = 4$ has the equation $= 1.32\ N^{-1.74}$ and the tangent at $N = 36$ has the ıuation $D = 2.21\ N^{-1.96}$. One would expect this sort behavior for large N. When N is large, the amplitude obability density does not vary appreciably from one ıd of a single input range to another, except for very rge amplitudes, which are sufficiently improbable so ıat their influence is slight. Hence, most of the output vels are very near to being the means of the end points the corresponding input ranges. Now, the best way quantizing a uniformly distributed input signal is to ace the output levels uniformly and to put the end oints of the input ranges halfway between the output vels, as in Fig. 2, shown for $N = 1$. The best way of roducing a quantizer with $2N$ output levels for this istribution is to divide each input range in half and ıt the new output levels at the midpoints of these ınges, as in Fig. 3. It is easy to see that the distortion the second case is $\frac{1}{4}$ that in the first. Hence, $D = kN^{-2}$ here k is some constant. In fact, k is the variance of ıe distribution.

If this sort of equal division process is performed on ıch input range of the optimum quantizer for a normally istributed signal with N output levels where N is large ıen again a reduction in distortion by a factor of 4 is xpected. Asymptotically then, the equation for the ıngent to the curve of distortion vs the number of output vels should be $D = kN^{-2}$ where k is some constant.

Commercial high-speed analog-to-digital conversion quipment is at present limited to transforming equal ıput ranges to outputs midway between the ends of ıe input ranges. In many applications one would like know the best interval length to use, *i.e.*, the one ielding minimum distortion for a given number of output vels, N. This is an easier problem than the first, since is only two-dimensional (for $N \geqq 2$), *i.e.*, D is a function of the common length r of the intervals and of ıy particular output level, y_k. If the input has a symmetric distribution and a symmetric answer is desired, ıe problem becomes one dimensional. If $p(x)$ is the input mplitude probability density and $f(x)$ is the function ıch that the distortion D is $E[f(s_{out} - s_{in})]$, then, for n even number $2N$ of outputs,

$$D = 2\sum_{i=1}^{N-1}\int_{(i-1)r}^{ir} f\left(x - \left[\frac{2i-1}{2}\right]r\right)p(x)\,dx + 2\int_{(N-1)r}^{\infty} f\left(x - \left[\frac{2N-1}{2}\right]r\right)p(x)\,dx. \tag{7}$$

For a minimum we require

$$\frac{dD}{dr} = -\sum_{i-1}^{N-1}(2i-1)\int_{(i-1)r}^{ir} f'\left(x - \left[\frac{2i-1}{2}\right]r\right)p(x)\,dx - (2N-1)\int_{(N-1)r}^{\infty} f'\left(x - \left[\frac{2N-1}{2}\right]r\right)p(x)\,dx = 0. \tag{8}$$

A similar expression exists for the case of an odd number of output levels. In either case the problem is quite susceptible to machine computation when $f(x)$, $p(x)$ and N are specified. Results have been obtained for $f(x) = x^2$, $p(x) = 1/\sqrt{2\pi}\, e^{-x^2/2}$, $N = 2$ to 36. They are indicated in Table II on page 12.

A log-log plot of distortion vs number of output levels appears in Fig. 1. This curve is not a straight line. The tangent to the curve at $N = 36$ has the equation $D = 1.47\ N^{-1.74}$. A log-log plot of output level spacing vs number of outputs for the equal spacing which yields lowest distortion is shown in Fig. 4. This curve is also not a straight line. Lastly, a plot of the ratio of the distortion for the optimum quantizer to that for the optimum equally spaced level quantizer can be seen in Fig. 5.

Key to the Tables

The numbering system for the table of output levels, y_i, and input interval end points, x_i, for the minimum mean-squared error quantization scheme for inputs with a normal amplitude probability density with standard deviation unity and mean zero is as follows:

> For the number of output levels, N, even, x_1 is the first end point of an input range to the right of the origin. An input between x_j and x_{j+1} produces an output y_j.
>
> For the number of output levels, N, odd, y_1 is the smallest non-negative output. An input between x_{j-1} and x_j produces an output y_j.

This description, illustrated in Fig. 6, is sufficient because of the symmetry of the quantizer. The expected squared error of the quantization process and informational entropy of the output of the quantizer are also tabulated for the optimal quantizers calculated.[2] (If p_k is the probability of the kth output, then the informational entropy is defined as $-\sum_{k=1}^{N} p_k \log_2 p_k$.)

Table II also pertains to a normally distributed input with standard deviation equal to unity. The meaning of the entries is self-explanatory.

[2] The values of informational entropy given show the minimum average number of binary digits required to code the quantizer output. It can be seen from the tables that this number is always a rather large fraction of $\log_2 N$, and in most cases quite near 0.9 $\log_2 N$. In the cases where $N = 2^n$, n an integer, a simple n binary digit code for the outputs of the quantizer makes near optimum use of the digital transmission capacity of the system.

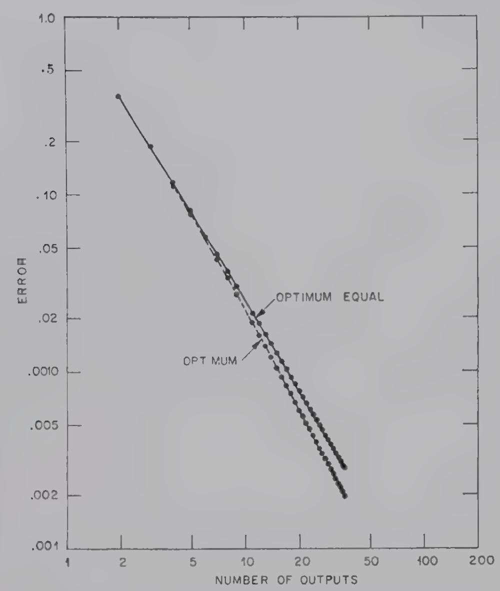

Fig. 1—Mean squared error vs number of outputs for optimum quantizer and optimum equally spaced level quantizer. (Minimum mean squared error for normally distributed input with $\sigma = 1$.)

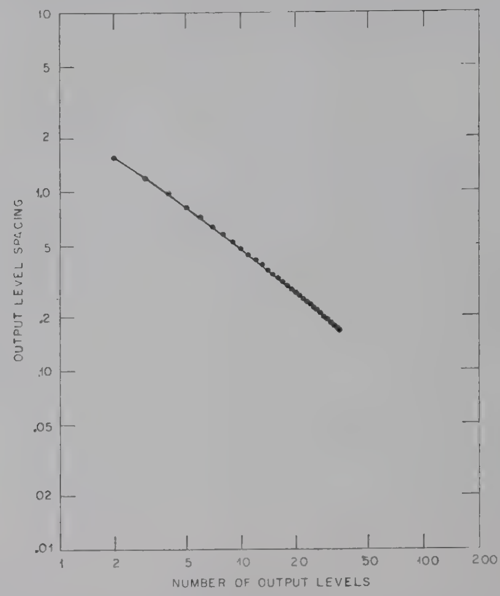

Fig. 4—Output level spacing vs number of output levels for equa optimum case. (Minimum mean squared error for normall distributed input with $\sigma = 1$.)

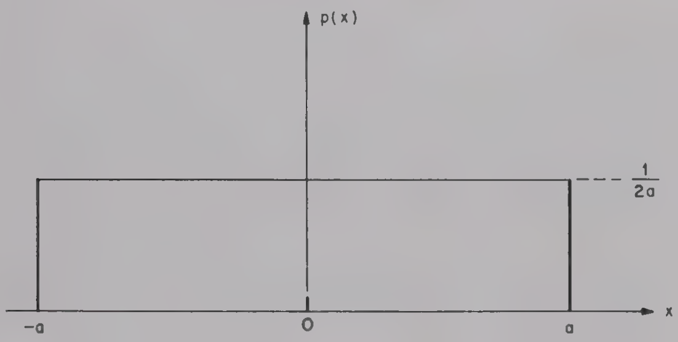

Fig. 2—Optimum quantization for the uniformly distributed case, $N = 1$. (Short strokes mark output levels and long strokes mark end points of corresponding input ranges.)

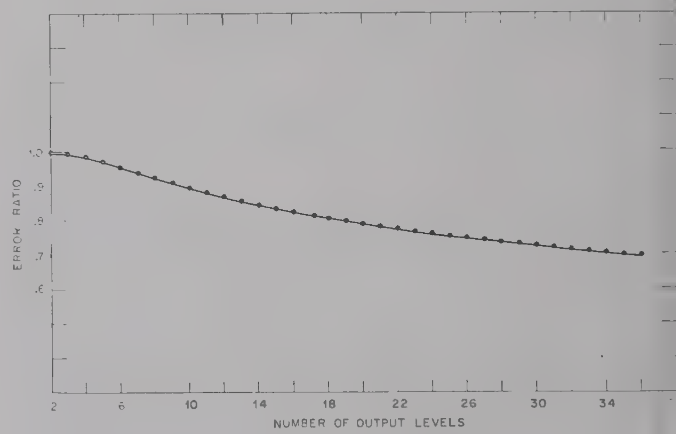

Fig. 5—Ratio of error for optimum quantizer to error for optimur equally spaced level quantizer vs number of outputs. (Minimur mean squared error for normally distribured input with σ = 1

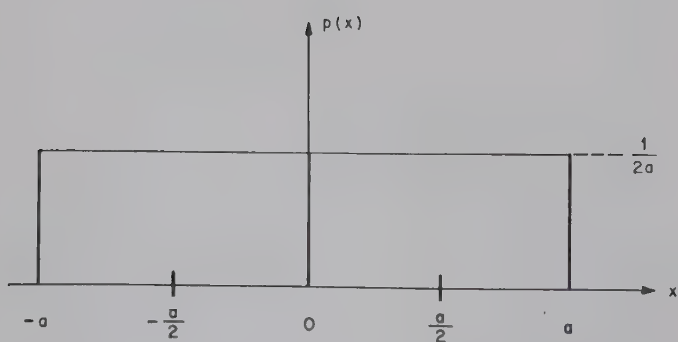

Fig. 3—Optimum quantization for the uniformly distributed case, $N = 2$. (Short strokes mark output levels and long strokes mark end points of corresponding input ranges.)

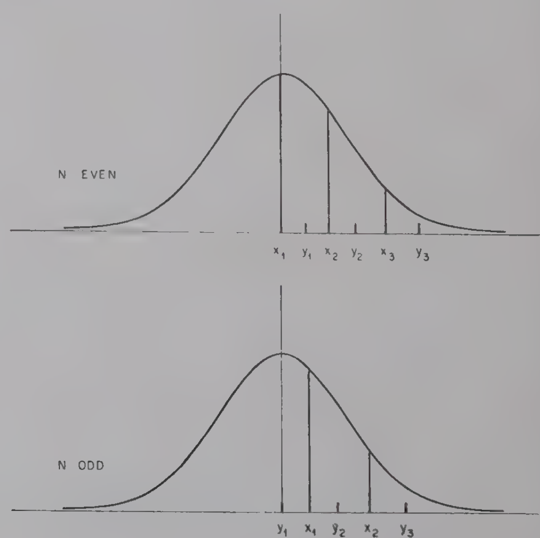

Fig. 6—Labeling of input range end points and output levels f the optimum quantizer. (Short strokes mark output levels a long strokes mark input range end points.)

TABLE I

Parameters for the Optimum Quantizer

	$N = 1$		$N = 2$		$N = 3$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	—	0.0	0.0	0.7980	0.0	1.224
2					0.6120	
Error	1.000		0.3634		0.1902	
Entropy	0.0		1.000		1.536	

	$N = 4$		$N = 5$		$N = 6$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.0	0.4528	0.3823	0.0	0.0	0.3177
2	0.9816	1.510	1.244	0.7646	0.6589	1.000
3				1.724	1.447	1.894
Error	0.1175		0.07994		0.05798	
Entropy	1.911		2.203		2.443	

	$N = 7$		$N = 8$		$N = 9$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.2803	0.0	0.0	0.2451	0.2218	0.0
2	0.8744	0.5606	0.5006	0.7560	0.6812	0.4436
3	1.611	1.188	1.050	1.344	1.198	0.9188
4		2.033	1.748	2.152	1.866	1.476
						2.255
Error	0.04400		0.03454		0.02785	
Entropy	2.647		2.825		2.983	

	$N = 10$		$N = 11$		$N = 12$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.0	0.1996	0.1837	0.0	0.0	0.1684
2	0.4047	0.6099	0.5599	0.3675	0.3401	0.5119
3	0.8339	1.058	0.9656	0.7524	0.6943	0.8768
4	1.325	1.591	1.436	1.179	1.081	1.286
5	1.968	2.345	2.059	1.693	1.534	1.783
6				2.426	2.141	2.499
Error	0.02293		0.01922		0.01634	
Entropy	3.125		3.253		3.372	

	$N = 13$		$N = 14$		$N = 15$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.1569	0.0	0.0	0.1457	0.1369	0.0
2	0.4760	0.3138	0.2935	0.4413	0.4143	0.2739
3	0.8126	0.6383	0.5959	0.7505	0.7030	0.5548
4	1.184	0.9870	0.9181	1.086	1.013	0.8512
5	1.623	1.381	1.277	1.468	1.361	1.175
6	2.215	1.865	1.703	1.939	1.776	1.546
7		2.565	2.282	2.625	2.344	2.007
8						2.681
Error	0.01406		0.01223		0.01073	
Entropy	3.481		3.582		3.677	

	$N = 16$		$N = 17$		$N = 18$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.0	0.1284	0.1215	0.0	0.0	0.1148
2	0.2582	0.3881	0.3670	0.2430	0.2306	0.3464
3	0.5224	0.6568	0.6201	0.4909	0.4653	0.5843
4	0.7996	0.9424	0.8875	0.7493	0.7091	0.8339
5	1.099	1.256	1.178	1.026	0.9680	1.102
6	1.437	1.618	1.508	1.331	1.251	1.400
7	1.844	2.069	1.906	1.685	1.573	1.746
8	2.401	2.733	2.454	2.127	1.964	2.181
				2.781	2.504	2.826
Error	0.009497		0.008463		0.007589	
Entropy	3.765		3.849		3.928	

	$N = 19$		$N = 20$		$N = 21$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.1092	0.0	0.0	0.1038	0.09918	0.0
2	0.3294	0.2184	0.2083	0.3128	0.2989	0.1984
3	0.5551	0.4404	0.4197	0.5265	0.5027	0.3994
4	0.7908	0.6698	0.6375	0.7486	0.7137	0.6059
5	1.042	0.9117	0.8661	0.9837	0.9361	0.8215
6	1.318	1.173	1.111	1.239	1.175	1.051
7	1.634	1.464	1.381	1.524	1.440	1.300
8	2.018	1.803	1.690	1.857	1.743	1.579
9	2.550	2.232	2.068	2.279	2.116	1.908
10		2.869	2.594	2.908	2.635	2.324
11						2.946
Error	0.006844		0.006203		0.005648	
Entropy	4.002		4.074		4.141	

	$N = 22$		$N = 23$		$N = 24$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.0	0.09469	0.09085	0.0	0.0	0.08708
2	0.1900	0.2852	0.2736	0.1817	0.1746	0.2621
3	0.3822	0.4793	0.4594	0.3654	0.3510	0.4399
4	0.5794	0.6795	0.6507	0.5534	0.5312	0.6224
5	0.7844	0.8893	0.8504	0.7481	0.7173	0.8122
6	1.001	1.113	1.062	0.9527	0.9122	1.012
7	1.235	1.357	1.291	1.172	1.119	1.227
8	1.495	1.632	1.546	1.411	1.344	1.462
9	1.793	1.955	1.841	1.681	1.595	1.728
10	2.160	2.366	2.203	2.000	1.885	2.042
11	2.674	2.982	2.711	2.406	2.243	2.444
12				3.016	2.746	3.048
Error	0.005165		0.004741		0.004367	
Entropy	4.206		4.268		4.327	

	$N = 25$		$N = 26$		$N = 27$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.08381	0.0	0.0	0.08060	0.07779	0.0
2	0.2522	0.1676	0.1616	0.2425	0.2340	0.1556
3	0.4231	0.3368	0.3245	0.4066	0.3921	0.3124
4	0.5982	0.5093	0.4905	0.5743	0.5537	0.4719
5	0.7797	0.6870	0.6610	0.7477	0.7202	0.6354
6	0.9702	0.8723	0.8383	0.9289	0.8936	0.8049
7	1.173	1.068	1.025	1.121	1.077	0.9824
8	1.394	1.279	1.224	1.328	1.273	1.171
9	1.641	1.510	1.442	1.556	1.487	1.374
10	1.927	1.772	1.685	1.814	1.727	1.599
11	2.281	2.083	1.968	2.121	2.006	1.854
12	2.779	2.480	2.318	2.514	2.352	2.158
13		3.079	2.811	3.109	2.842	2.547
14						3.137
Error	0.004036		0.003741		0.003477	
Entropy	4.384		4.439		4.491	

	$N = 28$		$N = 29$		$N = 30$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.0	0.07502	0.07257	0.0	0.0	0.07016
2	0.1503	0.2256	0.2182	0.1451	0.1406	0.2110
3	0.3018	0.3780	0.3655	0.2913	0.2821	0.3532
4	0.4556	0.5333	0.5154	0.4396	0.4255	0.4978
5	0.6132	0.6930	0.6693	0.5912	0.5719	0.6460
6	0.7760	0.8589	0.8287	0.7475	0.7225	0.7990
7	0.9460	1.033	0.9956	0.9100	0.8788	0.9586
8	1.126	1.218	1.172	1.081	1.043	1.127
9	1.319	1.419	1.362	1.263	1.217	1.306
10	1.529	1.640	1.570	1.461	1.404	1.501
11	1.766	1.892	1.804	1.680	1.609	1.717
12	2.042	2.193	2.077	1.929	1.840	1.964
13	2.385	2.578	2.417	2.226	2.111	2.258
14	2.871	3.164	2.899	2.609	2.448	2.638
15				3.190	2.926	3.215
Error	0.003240		0.003027		0.002834	
Entropy	4.542		4.591		4.639	

Cont'd next page

TABLE I, *Cont'd*

	$N = 31$		$N = 32$		$N = 33$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.06802	0.0	0.0	0.06590	0.06400	0.0
2	0.2045	0.1360	0.1320	0.1981	0.1924	0.1280
3	0.3422	0.2729	0.2648	0.3314	0.3218	0.2567
4	0.4822	0.4115	0.3991	0.4668	0.4530	0.3868
5	0.6254	0.5528	0.5359	0.6050	0.5869	0.5192
6	0.7730	0.6979	0.6761	0.7473	0.7245	0.6547
7	0.9265	0.8481	0.8210	0.8947	0.8667	0.7943
8	1.088	1.005	0.9718	1.049	1.015	0.9392
9	1.259	1.170	1.130	1.212	1.171	1.091
10	1.444	1.347	1.299	1.387	1.338	1.252
11	1.646	1.540	1.482	1.577	1.518	1.424
12	1.875	1.753	1.682	1.788	1.716	1.612
13	2.143	1.997	1.908	2.029	1.940	1.821
14	2.477	2.289	2.174	2.319	2.204	2.060
15	2.952	2.665	2.505	2.692	2.533	2.347
16		3.239	2.977	3.263	3.002	2.718
17						3.285
Error	0.002658		0.002499		0.002354	
Entropy	4.685		4.730		4.773	

	$N = 34$		$N = 35$		$N = 36$	
	x_j	y_j	x_j	y_j	x_j	y_j
$j = 1$	0.0	0.06212	0.06043	0.0	0.0	0.05876
2	0.1244	0.1867	0.1816	0.1209	0.1177	0.1765
3	0.2495	0.3122	0.3036	0.2423	0.2359	0.2952
4	0.3758	0.4394	0.4272	0.3650	0.3552	0.4152
5	0.5043	0.5691	0.5530	0.4895	0.4762	0.5372
6	0.6355	0.7020	0.6819	0.6166	0.5996	0.6620
7	0.7705	0.8391	0.8146	0.7471	0.7261	0.7903
8	0.9104	0.9818	0.9523	0.8820	0.8567	0.9231
9	1.057	1.131	1.096	1.023	0.9923	1.062
10	1.211	1.290	1.248	1.170	1.134	1.207
11	1.375	1.460	1.411	1.327	1.285	1.362
12	1.553	1.646	1.587	1.495	1.445	1.528
13	1.749	1.853	1.781	1.679	1.619	1.710
14	1.971	2.090	2.001	1.883	1.812	1.913
15	2.232	2.375	2.260	2.119	2.030	2.146
16	2.559	2.743	2.584	2.401	2.287	2.427
17	3.025	3.307	3.048	2.767	2.609	2.791
18				3.328	3.070	3.349
Error	0.002220		0.002097		0.001985	
Entropy	4.815		4.856		4.895	

TABLE II

PARAMETERS FOR THE OPTIMUM EQUALLY SPACED LEVEL QUANTIZER

Number Output Levels	Output Level Spacing	Mean Squared Error	Informational Entropy
1	—	1.000	0.0
2	1.596	0.3634	1.000
3	1.224	0.1902	1.536
4	0.9957	0.1188	1.904
5	0.8430	0.08218	2.183
6	0.7334	0.06065	2.409
7	0.6508	0.04686	2.598
8	0.5860	0.03744	2.761
9	0.5338	0.03069	2.904
10	0.4908	0.02568	3.032
11	0.4546	0.02185	3.148
12	0.4238	0.01885	3.253
13	0.3972	0.01645	3.350
14	0.3739	0.01450	3.440
15	0.3534	0.01289	3.524
16	0.3352	0.01154	3.602
17	0.3189	0.01040	3.676
18	0.3042	0.009430	3.746
19	0.2909	0.008594	3.811
20	0.2788	0.007869	3.874
21	0.2678	0.007235	3.933
22	0.2576	0.006678	3.990
23	0.2482	0.006185	4.045
24	0.2396	0.005747	4.097
25	0.2315	0.005355	4.146
26	0.2240	0.005004	4.194
27	0.2171	0.004687	4.241
28	0.2105	0.004401	4.285
29	0.2044	0.004141	4.328
30	0.1987	0.003905	4.370
31	0.1932	0.003688	4.410
32	0.1881	0.003490	4.449
33	0.1833	0.003308	4.487
34	0.1787	0.003141	4.524
35	0.1744	0.002986	4.560
36	0.1703	0.002843	4.594

A Note on *P*-nary Adjacent-Error-Correcting Codes*

BERNARD ELSPAS†

Summary—Binary group codes described by Abramson permit e correction of all single errors and all double errors in adjacent gits, with the use of significantly fewer check digits than codes pable of correcting all double-bit errors.

This note considers the generalization of Abramson's codes to e p-nary case, where a symbol alphabet consisting of the digits 1, $\cdots$, $p-1$ is used for transmission, p being a prime number. xamples of such p-nary codes are given, as well as necessary nditions for their existence. These codes bear the same relation the p-nary Golay codes as Abramson's codes do to the familiar amming codes.

Some as yet unanswered questions are raised, and suggestions r further possible generalizations are given.

I. Introduction

RECENT work by Abramson[1] has described the construction of binary group codes which are rather efficient for combatting burst noise. These single-ror-correcting double-adjacent-error-correcting codes EC-DAEC codes) use significantly fewer check digits ıan codes capable of correcting all double-bit errors. urthermore, the digital circuitry[2] required for imple-enting these codes is extremely simple.

This note describes a generalization of binary SEC-AEC codes to the p-nary case, where a symbol alphabet nsisting of the digits 0, 1, $\cdots$, $p-1$ is used for trans-ission, p being a prime number. These codes bear the me relation to the p-nary Golay codes[3] as Abramson's odes do to the Hamming SEC codes.

In the binary case, the block length n is related to e number of check digits k by the relation, $n = 2^{k-1} - 1$. his relation derives from the requirement that the ımber 2^k of corrector combinations possible with k ırity checks be at least $1 + 2n$ in order to cover all n ngle-bit errors and all n double-adjacent-bit errors, as ell as the case of no errors per block. Note that an error the first and last digits of a block is counted as a orrectible) double-adjacent error. Observe also that actly one k-bit combination remains unused for correc-on.

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[1] N. Abramson, "A Class of Systematic Codes for Non-Inde-endent Errors," Stanford Univ. Electronics Lab., Stanford, Calif., ept. No. 51; December 30, 1958. Also: IRE Trans. on Infor-ation Theory, vol. IT-5, pp. 150–157; December, 1959.

[2] N. Abramson and B. Elspas, "Double-Error-Correcting En-ders and Decoders for Non-Independent Binary Errors," pre-nted at the Internatl. Conf. on Information Processing, Paris, rance; June 15-22, 1959.

[3] M. J. E. Golay, "Notes on digital coding," Proc. IRE, vol. 37, 657; June, 1949.

II. The *P*-nary Case

In the p-nary generalization, the digit in any particular position may be in error in $p-1$ ways, and any pair of digit errors (in given locations) may occur in $(p-1)^2$ ways. Hence for a p-nary SEC-DAEC code we require that

$$1 + n(p-1) + n(p-1)^2 = 1 + np(p-1) \leq p^k. \qquad (1)$$

The left-hand side of the above inequality leaves a remainder of 1 upon division by p, while the right-hand side is divisible by p. It follows that integer solutions for n and k are constrained by

$$n \leq \frac{p^{k-1} - 1}{p - 1}. \qquad (2)$$

It is shown below that (at least for some values of k and p) SEC-DAEC codes exist for which the equality sign holds in the above condition. This relation between the block length and the number of check digits reduces to the binary formula when $p = 2$.

Any group code[4] is completely specified by a listing of the parity checks (modulo-p) which must be satisfied by a valid code word. Alternatively, this listing may be viewed as a specification of the n *characteristics* (*i.e.*, the correctors which result at the receiver when a single digit is received in error by an amount $+1$). Each characteristic, A_i, is a k-digit p-nary number. The corrector for an error of amount e occurring in the ith digit is then, by linearity of the parity checks, equal to eA_i. Likewise, the double-error corrector corresponding to an error of amount e in the ith digit and an error of amount f in the jth digit occurring simultaneously is then $eA_i + fA_j$. Thus if all single errors and all double-adjacent errors are to be correctible, the $np(p-1)$ correctors, $eA_i + fA_{i+1}$, (where $e = 1, 2, \cdots, p-1$; $f = 0, 1, \cdots, p-1$; $i = 1, 2, \cdots, n$ and all subscripts are taken modulo n), must all be distinct. Note that of the p^k conceivable correctors, $p-1$ remain unused for correction. The all-zeros corrector is, of course, reserved for the indication, "no error."

III. Examples of *P*-nary SEC-DAEC Codes

The simplest example of such a code is furnished by the ternary ($p = 3$) case for $k = 4$, $n = 13$. Here the thirteen required characteristics, A_i, are as shown in Fig. 1. The corresponding parity-check equations, in Fig. 2, are obtained by turning the list of characteristics

[4] D. Slepian, "A class of binary signalling alphabets," *Bell Sys. Tech. J.*, vol. 35, pp. 203–234; January, 1956.

Unit Error in Digit Number	Corrector
1	1001
2	0101
3	1011
4	1101
5	1111
6	2111
7	2211
8	0221
9	1021
10	2101
11	1211
12	0121
13	0011

Fig. 1—Correctors for ternary SEC-DAEC code with $n = 13$ and $k = 4$.

Data Digits — Check Digits

$$\begin{array}{ccccccccccccccccccccccccccc}
x_1 & + & . & + & x_3 & + & x_4 & + & x_5 & + & 2x_6 & + & 2x_7 & + & . & + & x_9 & + & 2y_1 & + & y_2 & + & . & + & . & = & 0 \\
. & + & x_2 & + & . & + & x_4 & + & x_5 & + & x_6 & + & 2x_7 & + & 2x_8 & + & . & + & y_1 & + & 2y_2 & + & y_3 & + & . & = & 0 \\
. & + & . & + & x_3 & + & . & + & x_5 & + & x_6 & + & x_7 & + & 2x_8 & + & 2x_9 & + & . & + & y_2 & + & 2y_3 & + & y_4 & = & 0 \\
x_1 & + & x_2 & + & x_3 & + & x_4 & + & x_5 & + & x_6 & + & x_7 & + & x_8 & + & x_9 & + & y_1 & + & y_2 & + & y_3 & + & y_4 & = & 0
\end{array}$$

Fig. 2—Parity check relations for the ternary SEC-DAEC code with $n = 13$ and $k = 4$.

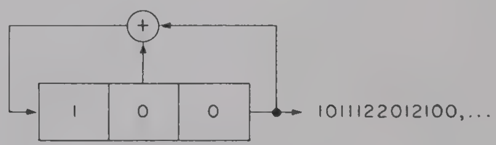

Fig. 3—Ternary feedback shift register with cycle length of 13.

on its side, yielding four parity checks over the thirteen code digits. One of these parity checks is simply a mod-3 sum of all the code digits, while the other three parity checks follow a particular pattern, cyclically permuted. Consideration of the A_i's shows that all of the 78 four-digit ternary numbers represented by A_i, $2A_i$, $A_i + A_{i+1}$, $2A_i + 2A_{i+1}$, $A_i + 2A_{i+1}$, $2A_i + A_{i+1}$ are distinct and nonzero, thus permitting the correction of the 26 possible single errors and the 52 possible double-adjacent errors.

A similar code (also ternary) has been found for $k = 6$, $n = 121$, which corrects all 242 possible single errors and all 484 possible double-adjacent errors in a block of 121 ternary digits by using six parity checks.

Both of the above codes were constructed by finding suitable feedback shift-register sequences to determine the cyclic parity check patterns. These shift-register sequences satisfy recursion relations of degree $k - 1$ (corresponding to shift registers with $k - 1$ stages), and their period (cycle length) is $n = (3^{k-1} - 1)/2$. For the code with $k = 4$, the recursion relation is

$$x_{i+3} = x_{i+1} + x_i \qquad (\text{mod } 3),$$

corresponding to the characteristic polynomial,

$$x^3 + 2x + 2 = 0,$$

which is irreducible over the ternary field (0, 1, 2), an which has period 13. The feedback shift-register employin this recursion relation is shown in Fig. 3.

The code for $k = 6$ can be constructed similarly i terms of the recursion relation,

$$x_{i+4} = x_{i+1} + x_i \qquad (\text{mod } 3)$$

which is equivalent to the characteristic polynomial,

$$x^4 + 2x + 2 = 0,$$

of period 121.

IV. Discussion

There is an extensive theory of feedback shift-registe and their characteristic polynomials; see for example th paper by Elspas.[5] This theory shows that the desir cycle length, n, as given by (2), can always be realiz for arbitrary k and p with several different shift-regist connections. However, it can be shown (see Appendi that two additional conditions must be met in order f these sequences to be applicable to the construction

[5] B. Elspas, "The theory of autonomous linear sequent networks," IRE Trans. on Circuit Theory, vol. CT-6, p. 45–6 March, 1959.

C-DAEC codes. The simpler of these conditions re-ires that n be relatively prime to $p - 1$. This condi-n has the effect of ruling out, *e.g.*, all even values of n e., odd values of k) for p (a prime) greater than 2. The other condition requires that if the characteristic lynomial, $f(x)$, divides (without remainder) a poly-mial of the form, $ax^{h+1} + bx^h - cx - d$, where $a + = c + d$, then $a = c$, and $b = d$. It is not presently iown for which values of p and k it is possible to find least one polynomial of degree $k - 1$ and period n at satisfies this condition, other than the two examples ed. However, it seems not unreasonable to suspect that ier cases exist.

The situation for $p = 2$ (the binary SEC-DAEC codes scribed by Abramson[1]) is much more clear cut since e two extra conditions are then automatically satisfied r all characteristic polynomials having the proper period, $= 2^{k-1} - 1$. Thus Abramson's codes exist (in shift-gister form) for all values of $k \geq 4$. When $k = 1, 2,$ 3, the situation is degenerate, since then $n \leq k$, and information can be transmitted.

The broader problem of the existence of SEC-DAEC des not derivable from shift-register sequences, is still open question (although one such case[1] is known for $= 2, n = 15$). A possible generalization to nonprime oduli (along the lines given by Cocke[6]) might also arrant investigation. Another possibly interesting gener-ization concerns the extension to p-nary codes of the ork of Fire[7] on binary burst-error codes where the irst length is greater than two.

V. Appendix

Necessary Conditions for the Existence of P-nary SEC-DAEC Codes

Theorem: In order for a p-nary shift-register having e (irreducible) characteristic polynomial $f(x)$ to be appli-ble to the construction of a p-nary SEC-DAEC code n digits with k check digits, the following conditions ust be satisfied:

1) Register cycle length $= n = (p^{k-1} - 1)/(p - 1)$ must be relatively prime to $p - 1$.

2) If $f(x)$ divides a polynomial of the form,

$$ax^{h+1} + bx^h - cx - d$$

where $a + b = c + d$, then $a = c$ and $b = d$ must hold.

Proof. It follows directly from (2), and the discussion of Section III, that the register cycle length must be given by

$$n = \frac{p^{k-1} - 1}{p - 1}.$$

Thus $f(x)$ divides $x^n - 1$.

1) If n is not relatively prime to $p - 1$, then

$$n = td \quad \text{and} \quad p - 1 = qd$$

for some integers, t and q. We now show that this implies that the resulting code based on $f(x)$ would leave an error of a certain amount in digit 1 indistinguishable from a unit error in digit $t + 1$. Thus this code would not even be single-error-correcting. We have:

$$x^n - 1 = x^{td} - 1 = y^d - 1, \quad \text{where} \quad y = x^t.$$

Now since d divides $p - 1$, $y^d - 1$ may be factored[8] into linear factors,

$$y^d - 1 = \prod_{i=1}^{d} (y - a_i) \qquad (\text{mod } p),$$

where the a_i are integers in the set $(1, 2, \cdots, p - 1)$. Since $f(x)$ divides $x^n - 1$, and is irreducible (mod p), it follows that $f(x)$ divides some factor $x^t - a_i$. But this in turn means that the shift-register sequence repeats (except for a constant factor a_i) in blocks of length t (a divisor of n). Such a sequence cannot be used for single-error correction of n digits since it leaves an error of amount a_i in digit 1 indistinguishable from a unit error in digit $t + 1$, for example. Hence n must be prime to $p - 1$.

2) If $f(x)$ divides $x^h(ax + b) - (cx + d)$ with $a + b = c + d$, then, as a little thought will show, errors of amounts d and c, respectively, in digits 1 and 2 will produce the same pattern of parity check violations as errors of amounts b and a, respectively, in digits $h + 1$ and $h + 2$. This means that the code is not adjacent-error-correcting, unless h is a multiple of n, and consequently, $a = c$ and $b = d$.

[6] J. Cocke, "Lossless symbol coding with nonprimes," Trans. E on Information Theory, vol. IT-5, pp. 33–34; March, 1959.

[7] P. Fire, "A Class of Multiple-Error-Correcting Binary Codes Non-Independent Errors," Sylvania Reconnaissance Sys. Lab., ountain View, Calif., Rept. No. RSL-E-2; March, 1959.

[8] See, for example, L. E. Dickson, "Introduction to the Theory of Numbers," Dover Publications, Inc., New York, N. Y., Theorem 21, p. 16; 1957.

Codes for the Correction of "Clustered" Errors*

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Summary—**A method is described which permits the systematic construction of codes capable of error-free transmission, provided errors occur in "clusters" of limited duration. The method is valid for error clusters of any prescribed duration. The codes are relatively easy to implement and decoding operations are straightforward. Specific examples are given and applications to teletype transmission are discussed.**

I. Introduction

THE construction of most error-correcting codes, which have been described in the literature, is based on the assumption that errors occur independently in digit positions. Representative codes in this category were described by Hamming,[1] Reed,[2] Slepian,[3] and Green and San Soucie.[4] With a digit error probability $p = 1 - q$, the usual assumption is that all possible s-tuple errors in a block of n digits are equally likely. (There are exactly $\binom{n}{s}$ s-tuple errors with total probability $\binom{n}{s}p^s q^{n-s}$.)

In many instances, errors do *not* occur independently. On telephone lines, for example, impulse noise is prevalent and high error probability is confined to the duration of the impulse, the error probability being negligible at other times. Similarly, on certain types of fading radio circuits, the signal is buried in noise for short periods of time. Ordinary "static," caused by thunderstorms, frequently displays such burst-like characteristics. Abramson[5] has recently described a class of codes which are useful in such cases. His codes correct all single and adjacent double errors and require significantly less checking digits than codes capable of correcting *all* double errors.

This paper describes a method which permits the construction of codes capable of correcting clustered errors of any prescribed cluster duration. The construction of these codes is simple and the decoding procedure is straightforward.

II. General Properties of Systematic Codes

This section introduces briefly some properties of systematic codes which are significant in the subsequent code construction. The notation is identical to Slepian's.[3]

* Manuscript received by the PGIT, August 7, 1959.
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[1] R. W. Hamming, "Error detecting and error correcting codes," *Bell Sys. Tech. J.*, vol. 29, pp. 147–160; April, 1950.
[2] I. S. Reed, "A class of multiple-error-correcting codes and the decoding scheme," IRE Trans. on Information Theory, vol. IT-4, pp. 38–49; September, 1954.
[3] D. Slepian, "A class of binary signaling alphabets," *Bell Sys. Tech. J.*, vol. 35, pp. 204–234; January, 1956.
[4] J. H. Green, Jr., and R. L. San Soucie, "An error-correcting encoder and decoder of high efficiency," Proc. IRE, vol. 46, pp. 1741–1744; October, 1958.
[5] N. M. Abramson, "A Class of Systematic Codes for Non-Independent-Errors," IRE Trans. on Information Theory, vol. IT-5, pp. 150–157; December, 1959.

In a systematic code, digit positions in any code wor of length n fall into two categories: information positions k in number, and $m = n - k$ check positions. Entrie in the check positions are linear (modulo 2) combina tions of information digits and are usually called parit checks. Without loss of generality, we may let the infor mation digits occupy the first k places of each code word Entries in the last $(n - k)$ positions are uniquely deter mined by parity checks over the first k digits. The parit check rules may be written

$$a_i = \sum_{j=1}^{k} \gamma_{ij} a_j \qquad i = k + 1, \cdots, n \qquad (1$$

which is Slepian's (9). In (1), a_i is the digit in position and is 0 or 1. Additions are modulo 2, *viz.*, $1 + 1 = 0$ - $0 = 0$, $0 + 1 = 1 + 0 = 1$. Eq. (1) defines $m = n -$ linear relations, one for each check digit.

The key problem in the design of a code having specifie optimal properties is the determination of the matrix γ_i No straightforward, general rules have as yet been foun to accomplish this.

We modify (1) slightly for reasons which will becom apparent later. Each check digit will be allowed to perforr a parity check not only over information digits bu also over check digits in preceding places. Thus, th second check digit, a_{k+2}, may also check a_{k+1}, etc. Th check rules now read

$$a_i = \sum_{j=1}^{i-1} \alpha_{ij} a_j \qquad i = k + 1, \cdots, n \qquad (2$$

which is similar to (1) except that the summation no goes from $j = 1$ to $i - 1$. With $\alpha_{ii} = 1$, we may writ

$$0 = \sum_{j=1}^{i} \alpha_{ij} a_j \qquad i = k + 1, \cdots, n. \qquad (2a$$

In encoding, each check digit can still be written as linear combination of information digits only, becaus a_{k+1} depends only on information digits and can b substituted by a linear combination of information digit in the equation for a_{k+2}, and so on. The code remair systematic as defined above. The distinction between (1 and (2) lies in the decoding process.

Let b_i be the digits after reception, $i = 1, 2 \cdots$ If an error occurred in position i, we have $b_i \neq a_i$, $a_i + b_i = 1$ because of the rules for modulo 2 additio If the digit in position i was received correctly, $b_i =$ or $a_i + b_i = 0$. It is possible, then, to introduce an place binary number, which we might call error vecto the ith component of which is $e_i = a_i + b_i = 1$, if error occurred in position i and $e_i = 0$ otherwise. Th number of ones in the error vector is identical to t

umber of transmission errors. Also, any combination of rrors is defined uniquely by an error vector.

As a first step in decoding, the decoder performs $\imath = n - k$ summations (modulo 2) as prescribed by he right-hand side of (2a); in other words, m digits are omputed as follows:

$$r_q = \sum_{j=1}^{i} \alpha_{ij} b_j \qquad i = k + 1, \cdots, n \qquad (3)$$

$$\alpha_{ii} = 1, \qquad \alpha_{ij} = 0 \quad \text{for} \quad j > i$$

$$q = i - k = 1, 2, \cdots m.$$

Ve may interpret the m values of r_q as digits of an m-place inary number R, which we call *Check Number*. (Slepian alls it parity check sequence.) The modulo 2 sum of 2a) and (3) is

$$r_q = \sum_{j=1}^{i} \alpha_{ij}(a_j + b_j) = \sum_{j=1}^{i} \alpha_{ij} e_j. \qquad (4)$$

'he check number for error-free transmission is the m-lace zero sequence. In addition to the all-zero sequence, ' may possess any one of $(2^m - 1)$ values. Hence, a code in correct a maximum of $(2^m - 1)$ distinct error combiations, which is possible only if an α_{ij} matrix can be ound such that the $(2^m - 1)$ distinct error vectors (for rror combinations to be corrected) when introduced into 1), generate $(2^m - 1)$ *distinct* check numbers. In this ase the code is called "complete." In general, codes are complete, because check numbers are interrelated and ust satisfy additional conditions.

Consider the occurrence of a single error in position t id designate the associated check number by $R_t(1)$. nce a single error may occur in any one of n positions, iere are n check numbers $R_t(1)$. The first digit of $R_t(1)$ $r_1 = \alpha_{k+1,t}$, which follows from (4) with $e_t = 1$, and $= 0, j \neq t$. If the first check digit checks position t, e have $\alpha_{k+1,t} = 1$, because of (2), and $R_t(1)$ has a 1 the first position. In general, if $R_t(1)$ has a 1 in position parity check q checks position t and $\alpha_{k+q,t} = 1$. Thus, $_t(1)$ is column t of the a_{ij} matrix. We will subsequently fer to this fact as Rule I. Because of this one-to-one rrespondence between the α_{ij} matrix and the array single-error check numbers $R_t(1)$, a systematic code ay be described completely by either one.[6] The check imber concept is more useful in case of clustered errors.

We now turn to relations between check numbers. iere are two reasons for such relations; first, the linear rm of (2), (3) and (4), and second, the systematic ture of codes. An error vector, as defined above, idenying simultaneous errors in several positions is the nary sum of error vectors belonging to single errors those positions. Because of the linear form of (4), is property also holds for check numbers. Therefore, *check number for any multiple error is the linear (modulo 2) sum of check numbers $R_t(1)$ for single errors in the error positions of the multiple error.* We shall refer to this property as Rule II.

The multitude of linear relations between check numbers is the reason why codes are, in general, incomplete. Additional restrictions are placed on the choice of check numbers by the nature of systematic codes. Imagine the portion of the $R_t(1)$ array identifying single errors in the last m positions of the code word. Because of $\alpha_{ii} = 1$, the first of these numbers must have a 1 in the first position, the second in the second position, etc. Furthermore, since the first parity check may only include preceding positions, all numbers in this partial array except the first must have zeros in the first position, etc., because of Rule I. For example, in case of 5 check digits, the structure of the last five numbers in the array $R_t(1)$ must be as follows:

$q =$	1	2	3	4	5
R_{k+1} (1)	1	*	*	*	*
R_{k+2} (1)	0	1	*	*	*
R_{k+3} (1)	0	0	1	*	*
R_{k+4} (1)	0	0	0	1	*
R_{k+5} (1)	0	0	0	0	1.

(5)

Entries in the starred places are still at our disposal as well as the entire check numbers belonging to the first k positions of the code word, subject to the restrictions imposed by Rule II and the condition that check numbers for correctable error combinations must all be different from each other.

III. Check Numbers for Clustered Errors

The discussion in the previous section is quite general and applies to any systematic code governed by check rules as in (2). Consider now situations in which errors occur in clusters. The characteristic feature is the occurrence of a high error probability for the duration of s successive digit positions. The error probability is negligible in the remaining $(n - s)$ positions. Let us designate codes which are capable of effecting error-free transmission under these conditions as $(n, k)_s$ codes.

Up to length s, any combination of errors and correct digits is likely. Let h be the number of digit positions bracketed by the first and last error of an error combination, these two positions included. There are 2^{h-2} distinct error patterns of length h because errors may or may not occur in $(h - 2)$ places between the first and last error position. Error patterns of length h may be identified by h-place binary numbers $(1 \cdots 1)$ with a 1 denoting an error and a 0 no error. For example, if $h = 4$, there are $2^{h-2} = 2^2 = 4$ distinct error patterns, in this notation (1001), (1101), (1011) and (1111). The total number of multiple error patterns of an $(n, k)_s$ code is obtained by summation over h from 2 to s:

$$\sum_{h=2}^{s} 2^{h-2} = 2^{s-1} - 1. \qquad (6)$$

[6] G. E. Sachs, "Multiple error correction by means of parity ecks," IRE Trans. on Information Theory, vol. IT-4, pp. 5–147; December, 1958.

Associated with these patterns are check numbers, which we write $R_t(1 \cdots 1)$, where the subscript t identifies the position of the last error. For example, in case $s = 4$, we obtain $2^3 - 1 = 7$ check numbers with last error in position t, namely $R_t(11)$, $R_t(101)$, $R_t(111)$, $R_t(1001)$, $R_t(1011)$, $R_t(1101)$, $R_t(1111)$. As a consequence of Rule II, all of these are linear combinations of $R_t(1)$, $R_{t-1}(1)$, $R_{t-2}(1)$ and $R_{t-3}(1)$.

As the position of the last error is shifted through the code word, $2^{s-1} - 1$ distinct sequences of multiple error check numbers occur, one for each error pattern. Each of these sequences contains $(n + 1 - h)$ members, because $(n + 1 - h)$ is the number of distinct locations of an error pattern of length h in a code word of length n, $2 \leq h \leq s$. In addition, there is the sequence of n single error check numbers $R_t(1)$ plus the all zero m-place number for correct transmission. Therefore, the total number of distinct check numbers required for an $(n, k)_s$ code is

$$N(s) = 1 + n + \sum_{h=2}^{s} (n + 1 - h)2^{h-2} = (n + 2 - s)2^{s-1}. \quad (7)$$

This defines a lower bound for m, the necessary number of redundant digits:

$$2^m \geq (n + 2 - s)2^{s-2} \quad (8)$$

where the equality would hold for a "complete" code. These relations have also been obtained by Fire.[7]

A further restriction on m arises as follows: suppose an error cluster of length s happens somewhere within the last m positions. The code has to cope with 2^s possible errors vectors, which must be uniquely identifiable by the remaining $(m - s)$ redundant digits. Therefore

$$m \geq 2s. \quad (9)$$

For example, regardless of word length n, no $(n, k)_5$ code (cluster length 5) can be found which has less than 10 (redundant) check digits.

A simple consideration shows that half of the error combinations contain an odd number of errors, if zero-error is included in the count for even numbers of errors.

IV. Generation of Check Numbers with Linear Shift Registers

It has been demonstrated that $(n, k)^s$ codes are completely defined by a set of n single error check numbers $R_t(1)$. Furthermore, several conditions have been discussed, which must be met by the check numbers. As usual, there is no hint as to how to select an optimum set of check numbers. In principle, one could select the last m numbers in accordance with array (5) adjusting entries in the starred positions such that conflicts of multiple error check numbers do not occur. One migh continue to select the remaining check numbers, on at a time, rechecking the entire array after each step Clearly, such a procedure would result in chaos for larg n and s. In addition, entirely untractable decoding rule would be the inevitable consequence.

Linear feedback shift-registers offer a systematic metho of generating sequences which may be quite easily adapte for use as check numbers. We shall see that codes con structed by this method are quite efficient, though gener ally incomplete. Moreover, the decoding process is straight forward and simple.

A linear feedback shift-register is shown in Fig.

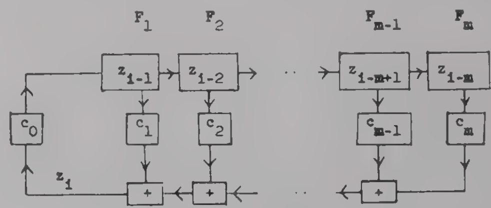

Fig. 1—Linear feedback shift-register.

(Their properties have been studied primarily by Zierle and Golomb.[9]) An m-stage shift register consists of m binary memory devices (flip-flops $F_1, \cdots F_m$), m switche $c_0, c_1, \cdots c_m$, which are either open or closed, and $(m - 1)$ binary adders. Imagine the device driven by clock pulse each of which shifts the flip-flop contents one positio to the right. Simultaneously, an input z_i is entered int the first stage where z_i is given by

$$z_i = \sum_{j=1}^{m} c_j z_{i-j}, \quad \text{or} \quad 0 = \sum_{j=0}^{m} c_j z_{i-j} \quad (10)$$

and where $c_0 \equiv c_m \equiv 1$.

The following properties of sequences z_i are of signi ficance here:

1) Sequences are periodic. Certain sequence generato $C = (c_0 c_1 \cdots c_m)$ generate sequences of peric $(2^m - 1)$, called maximum-length sequences. Max mum-length sequence generators have been tabu lated up to $m = 19$ by Marsh[10] and are also liste up to length 9 in Albert.[11] It is convenient to chara terize generators in octal equivalents of three-plac binary numbers, *viz.*, $0 = 000$, $1 = 001$, $2 = 01$ $3 = 011$, $4 = 100$, $5 = 101$, $6 = 110$, $7 = 11$ For example, the generator $1423 - (00110001001$ identifies a 9-stage shift register in which the su

[7] P. Fire, "A Class of Multiple-Error Correcting Binary Codes for Non-Independent-Errors," Stanford Electronics Labs., Stanford University, Stanford Calif., Tech. Rept. No. 55, Contract Nonr 225(24), NR 373 360; April, 1959.

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[9] S. W. Golomb, "Sequences with Randomness Propertie Glenn L. Martin Co., Baltimore, Md., Final Rept. on Contract N W36-039 Sc-54-3611; June 14, 1955.

[10] R. W. Marsh, "Table of Irreducible Polynomials Over GF (Through Degree 19," Natl. Security Agency, Washington, D. C October, 1957.

[11] A. A. Albert, "Fundamental Concepts of Higher Algebra University of Chicago Press, Chicago, Ill.; 1957.

of the contents of stages 1, 5, 8, and 9 is fed back to the input. We shall call a maximum-length sequence an m sequence.

2) There are exactly $\phi(2^m - 1)/m$ distinct m sequences of length $2^m - 1$. $\phi(2^m - 1)$ is the Euler function, equal to the number of integers (1 included) smaller than $(2^m - 1)$ and coprime (no common factor except 1) with $(2^m - 1)$.
3) The reverse of each m sequence has also period $(2^m - 1)$. The corresponding pair of sequence generators C^* and C is related by $c_i^* = c_{m-i}$.
4) An m sequence contains one run of m consecutive ones and one run of $(m - 1)$ zeros. Also a run of $(m - 2)$ ones, as well as a run of $(m - 2)$ zeros occurs exactly once. In general, there are 2^i runs of ones and 2^i runs of zeros of length $(m - i - 2)$, for $0 \leq i \leq m - 3$. A run of m zeros does not occur.
5) Consider the digits in each m consecutive positions as an m-place binary number. The result is a sequence of $(2^m - 1)$ m-place binary numbers. Each occurs exactly once, with the exception of the m-place all-zero number. We will call this sequence the M sequence.

In addition to the properties listed above, m sequences ssess the so-called "cycle and add" property which, fact, is their outstanding feature for our purposes. ne designation cycle and add means that a new sequence tained by adding termwise (modulo 2) an m sequence a cyclic permutation of itself is yet another cyclic rmutation of the original sequence. For example, if add consecutive digit pairs, we obtain the original sequence shifted by a certain number of positions. epeating the process, we see that the property holds r the sum of any digit pattern. Still more important, e cycle and add property also holds for the M sequence nply because it holds for each individual position in e m-place numbers of the M sequence.

Recall now Rule II concerning check numbers. Suppose take as single error check numbers, $R_t(1)$, a run of n nsecutive numbers from an M sequence. The cycle and d property then insures that the set of $(n + 1 - h)$ eck numbers for a given error pattern of length h is run of $(n + 1 - h)$ consecutive numbers in the M quence, and this is true for all error patterns, because ultiple error check numbers are linear sums of single ror check numbers (Rule II). These runs of consecutive mbers are shifted with respect to each other in the sequence. Unfortunately, there is no simple explicit rmula giving the amount of the various shifts as a nction of the sequence generator C. The realizable code rd length n, given a cluster length s, and a number of eck digits m, depends only on C. The limitation is at no two check numbers may be identical. Identity check numbers would occur if any two runs of check mbers overlap in the M sequence. In order to determine the maximum code length, it is only necessary to d the position of the first number in each run, and it is quite immaterial where in the M sequence we place the run of $R_t(1)$. A more detailed discussion of this procedure is found in the Appendix.

The actual check numbers for a realizable code can be chosen in one of two ways. In the first method, the m-place check numbers are taken from an M sequence. Because of array (5) we have to place the run of $R_t(1)$ within the M sequence so as to make $R_n(1)$ that number which has zeros in the first $(m - 1)$ places and a one in the last place. Thus, all check numbers are fixed and the check rules are found with Rule I.

It becomes clear now, why (2) had to be used instead of (1) to define the parity checks. Eq. (1) would require that all starred positions in array (5) be zeros; thus, a string of $(m - 1)$ zeros would be needed in $R_n(1)$ and $R_{k+1}(1)$. There is only one run of $(m - 1)$ zeros in an m sequence. Hence, parity check rules as in (1) are incompatible with the properties of m sequences.

An alternate method of check number assignment is based on the previously stated fact that half the error combinations contain an even number of errors (zero errors included), the other half containing an odd number of errors. This suggests a separation of these two groups by an over-all parity check to be performed by the last check digit in position n. Check numbers for even errors, then, have a zero in the last place and check numbers for odd errors have a one in the last place. The remaining $(m - 1)$ positions (truncated check numbers) may now be taken from $(M - 1)$ sequences. A slight complication arises, because an $(M - 1)$ sequence does not contain the $(m - 1)$-place all-zero number. But array (5) shows that this is the truncated check number required for $R_n(1)$. This difficulty may be resolved by an exchange of zeros and ones in the $(M - 1)$ sequence, because the all-one number does occur. This leaves the check numbers for even errors unaltered, but exchanges zeros and ones in the odd-error check numbers. No additional restrictions are imposed, however, because the all-parity check separates the two cases. Only conflicts in the $(M - 1)$ sequence between runs of truncated check numbers for even errors and conflicts between odd errors restrict the code length.

It is not now known how to predict *a priori* which of the two methods, shift register of length m or all-parity-check and shift register of length $(m - 1)$ produces longer codes; it depends entirely on the structure of the sequences.

V. Decoding

Decoding is particularly straightforward. Parity checks are performed on the received message in accordance with (3) and the resulting check number is stored. Check numbers are generated by a maximum length shift register in initial state $R_1(1)$. A counter, counting the shift pulses, is reset every time a new run of check numbers starts. When coincidence occurs between the received check number and the locally generated check number,

the number of resets of the counter identifies the error pattern, and the number in the counter identifies its position. The same procedure can be used with the all-parity-check codes, by running through the cycle twice, once for even numbers of errors and once for odd numbers.

Because of the logical simplicity of the decoder, a more detailed decoder design is omitted.

VI. Realizable Codes

Abramson[5] has constructed codes for the correction of single and adjacent double errors. We are, therefore, interested in codes which correct errors in at least three adjacent positions. This means $m \geq 6$. All maximum length sequences generated by shift registers of length 5, 6, 7, 8, and 9 have been investigated manually. As a result, codes were obtained of both types described above, m-place check numbers generated by m-stage shift registers and m-place check numbers generated by $(m - 1)$ stage shift registers plus all-parity check. Table I lists the longest realizable codes together with their sequence generators. Note that in each instance the reverse sequence generates a code of same length. Sequence generators for codes containing an all-parity check are underlined. (See Table I.) The Table can be extended without difficulty, preferably with the help of a digital computer.

TABLE I
Realizable $(n, k)_s$ Codes

	$s = 3$		$s = 4$		$s = 5$	
	C	n	C	n	C	n
$m = 6$	$\underline{45}$	11	—	—	—	—
$m = 7$	$\underline{147}$	26	—	—	—	—
	211	26	—	—	—	—
$m = 8$	$\underline{313}$	40	$\underline{211}$	15	—	—
	543	33	435	18	—	—
$m = 9$	$\underline{607}$	100	$\underline{543}$	34	—	—
	1423	114	1131	38	—	—
$m = 10$	$\underline{1055}$	206	$\underline{1131}$	64	$\underline{1021}$	19

Note: Sequence generators for codes with all-parity check are underlined.

The Table indicates that for each added check digit, for a given s, the maximum code word length is roughly doubled, as one would expect. It is not clear, however, which of the two described assignment methods is better in general.

Finally, a word about the efficiency of the codes which might be measured in terms of the length of hypothetical complete codes as defined by the equality sign in (8) is in order. Given m and s, complete codes would permit word lengths $n_C = 2^{m+1-s} - 2 + s$, which follows fro (8). For $s = 3$, the tabulated codes have values n/n ranging from 61 to 88 per cent and for $s = 4$, 48 to 5 per cent. For any value $n/n_C > \frac{1}{2}$, no code can exi which will allow the tabulated word length and wi correct clusters up to s, but has less redundant digi than the listed code. Most codes in columns $s = 3$ an $s = 4$ meet this criterion. Since only shift registers u to length 9 were investigated, the single code (19, 9 which was obtained, tells little about $s > 4$.

VII. Example

The following example of the construction of a co will illustrate the method. The first code listed in Table is 11 digits along with six check digits. The sequen generator is listed as 45 — (100101), a 5-stage shi register. The sum of the third and fifth stage are fe back. We start the sequence with 5 ones, which for th truncated check numbers are then changed into fi zeros. The $(m - 1)$ sequence and the position of th truncated check numbers of error patterns ending i position 11 are shown below. (Error patterns are show with each number.)

$$\underbrace{1\,1\,1\,1\,1}_{(1)}\,\underbrace{0\,0\,0\,1\,1}_{(101)}\,0\,\underbrace{1\,1\,1\,0\,1}_{(111)}\,0\,1\,\underbrace{0\,0\,0\,0\,1}_{(11)}\,0\,0\,1\,0\,1\,1\,0$$

There are $2^{s-1} = 4$ distinct runs of check number Since the code has an all-parity check, only conflic between even numbers of errors and conflicts betwee odd numbers of errors have to be considered. Note th $R_i(1)$ shifted by 11 positions is in conflict with the chec number for triple errors. Thus, the maximum code lengt is $n = 11$. The two runs for double errors (11) and (10 start 13 positions apart, which is no further restrictio Starting with the truncated check numbers shown i the sequence above, we may obtain 2 different cod of same length by continuing the check number ru either to the right or to the left. Fig. 2, shows the chec rules (α_{ij}) for the $(11, 5)_3$ code obtained by continuin the check number runs to the left.

	Inf. Digits					Check Digits					
i \ j	1	2	3	4	5	6	7	8	9	10	11
6		1			1	1					
7	1		1			1	1				
8	1	1		1			1	1			
9		1	1		1			1	1		
10	1		1	1		1			1	1	
11	1	1	1	1	1	1	1	1	1	1	1

Fig. 2—Check rules for $(11, 5)_3$ code.

The code is capable of correcting any errors whi occur in any 3 adjacent positions. The construction longer codes and for $s > 3$ is equivalent.

VIII. Application

Consider the application of $(n, k)_s$ codes to a common rm of digital communication, namely teletype. The iestion is the maximum duration of a noise burst with hich a code can cope successfully. The information in groups of five digits. Let us measure time in units length of the original information digits (for example .4 msec at a nominal speed of 60 wpm). Assume that ie receiver will receive a digit correctly if normal recep- on takes place during at least half the duration of the git. This assumption, although somewhat arbitrary, is asonable in many practical cases. Since the noise burst, gnal fade, or whatever else causes the disturbance curs at random times with respect to a digit, a disturb- ice of duration between one-half and one and one- lf times a digit duration may cause a high error prob- ility for a single digit position. Also, no disturbance length less than one digit length can corrupt two ljacent digits. Hence a code, which is capable of cor- cting errors in s adjacent position can cope with any sturbance up to a duration of s digits no matter what le relative timing is.

Consider first the encipherment of individual teletype iaracters. To be realistic, we must assume that we ish to maintain a constant information rate. As a con- quence, the digit durations after encoding are not unity for uncoded transmission but k/n, and the permissible irst length is sk/n. Table II lists the maximum per- issible duration of a disturbance for the codes described this paper, together with corresponding numbers for me codes, which have been published previously.

TABLE II
Permissible Duration T of Disturbance

Type of Code	n	T	Remarks
o code	5	0.5	
amming[1]	9	0.55	Detects double errors
oramson[5]	10	1.00	Corrects adjacent double errors
eed[2]	16	0.94	Correct *all* triple errors
reen-Sans Soucie[4]	15	1.00	
$1, 5)_3$	11	1.36	$s = 3$
$3, 5)_4$	13	1.54	$s = 4$
$5, 5)_5$	15	1.67	$s = 5$

If disturbances do not occur too frequently, it is ad- ntageous to encode a group of several teletype characters to one block. Table I suggests these possibilities:

	Code	Disturbance length
1) 2 characters	$(18, 10)_4$	2.22
2) 5 characters	$(34, 25)_4$	2.94
3) 10 characters	$(60, 50)_4$	3.33

Thus, at a nominal 60-wpm speed, a code encoding, for example, five teletype characters into one block of length 34, can handle disturbances of duration up to about 100 msec, provided such disturbances do not occur more often than about once per second. In the process, errors in up to four adjacent positions are corrected. We might also note that a code for five teletype characters capable of correcting up to *all* quadruple errors must have at least 18 check digits, instead of nine in our $(34, 25)_4$ code, and the maximum disturbance length would be reduced to 78 msec.

Appendix: Determination of Maximum Code Length

Assume for the moment that the run of consecutive numbers $R_t(1)$ starts with the all-one number in the M sequence and call this assignment $R_t(1)^*$. For convenience, we take $R_n(1)^* = (111..11)$ and to a single error in position $n - 1$ we assign the next number of the M sequence which has ones in the first $(m - 1)$ places and a zero in the last place. The first number of the run $R_t(11)^*$ has then zeros in the first $(m - 1)$ places and a one in the last place, because $R_n(11) = R_n(1) + R_{n-1}(1)$. In general $R_n(1 \cdots 1)^*$ has zeros in the first $(m + 1 - h)$ places if the associated error pattern contains an even number of errors. Conversely, for odd numbers of errors $R_n(1 \cdots 1)^*$ starts with $(m + 1 - h)$ ones. It follows that the numbers $R_n(1 \cdots 1)^*$ are in those positions of the M sequence where the runs of zeros and ones, described in property 4 of m sequences (see Section IV) represent the first $(m + 1 - h)$ places of corresponding m-place binary numbers of the M sequence.

There are 2^u runs of length $(m - u - 1)$ of ones *or* zeros, and 2^{h-2} runs of check numbers, $2 \leq h \leq s$. Putting $u = h - 2$, we see that each run of $(m + 1 - h)$ ones or zeros represents the first $(m + 1 - h)$ places of a run of $R_n(1 \cdots 1)^*$. Thus, in order to determine n for an $(n, k)_s$ code, it is only necessary to inspect in the m-sequence runs of zeros and ones longer than $(m - s)$ and count the number of shifts between these runs. Let n' be the number of shifts between two such runs. That is, with $2 \leq h \leq s$ and $2 \leq h' \leq s$, assume that the first zero or one of a run of $(m + 1 - h)$ zeros or ones in the m sequence, when shifted by n' positions, coincides with the first zero or one of a run of length $(m + 1 - h')$. A run of length $(m + 1 - h)$ belongs to an error pattern of length h, which has $(n + 1 - h)$ check numbers. Therefore, $(n + 1 - h) = n'_{\min}$ or $n = n'_{\min} + h - 1$. When there are several numbers n' nearly equal with different h, this must clearly be refined to read $n = (n' + h)_{\min} - 1$.

A Class of Codes for Signaling on a Noisy Continuous Channel*

JOHN L. KELLY, JR.†

Summary—**A class of codes for continuous channels is described. These are block codes in which the code words can be computed from a much smaller set of generators. It is shown that codes of this type exist which will yield arbitrarily small error rates at any signaling rate below channel capacity. In fact, if the generators are chosen at random, it is shown that the expected error rate obeys a bound established by Shannon for general random codes.**

Introduction

It is well known that, if a block code (continuous or discrete) is generated by choosing the code words at random with the proper distribution and if the signaling rate is kept below channel capacity, the expected probability of error will approach zero with increasing code length.[1] It is likely, however, that codes generated in this manner would have so little structure that the entire list of code words would have to be stored at the transmitter. It would be desirable for the transmitter to be able to compute the code words from some sort of algorithm.

A class of codes called "Additive codes" is described below in which 2^{NR} code words of length N are computed from $NR + 1$ generators, where R is the signaling rate in bits per unit time. These additive codes for continuous channels are somewhat analogous to group codes for binary channels in which 2^{NR} code words are computed from NR generators in a linear fashion. It will be shown that if the generators for an additive code are chosen at random, the expected probability of error obeys a bound given by Shannon[2] for general random codes. Also the probability is high for large N that the code words will meet the channel constraints.

Notation and Formalism

We avoid mathematical difficulties by considering channels whose inputs and outputs are sequences of real numbers occurring at discrete times. With certain reservations, each sequence may be thought of as the sample values of a band-limited time function. For N fixed but arbitrary, let

$$X = (X_1 \cdots X_N); \qquad Y = (Y_1 \cdots Y_N)$$

where the X_r and Y_r are random variables representing the input and output samples, respectively, of a noisy communication channel. X and Y are thus rando vectors. Let

$$x = (x_1 \cdots x_N); \qquad y = (y_1 \cdots y_N)$$

where the x_r and y_r are real numbers. In general, low case letters will stand for numbers and capitals for rando variables. The channel noise is described by a set conditional distributions (one for each N) of Y cond tional on X. We will consider only those cases whe the noise is represented by density functions $P(y/$ such that:

$$\text{probability} \quad [Y \varepsilon A/X = x] = \int_A P(y/x)\, dy.$$

We state the channel constraint in the following for to cover cases where the input stochastic process is no stationary (as will be the case with block codes)

$$\lim_{T\to\infty} \frac{1}{T} \sum_1^T f(X_r) < f_{\max} \qquad (1$$

with probability one where $f(\cdot)$ is a positive functio and $f_{\max}$ is a number. For example, if $f(X) = X^2$, w have an average power limitation.

By a "code of length N and signaling rate R" is mea a set of 2^{NR} labeled vectors $x^1 \cdots x^m$, m $= 2^{NR}$ wit

$$x^r = (x_1^r, x_2^r, \cdots x_N^r).$$

The x^r are called code words.

By a source is meant a method of assigning probabil ties to the stochastic process, X. We will consider tw sources. One, called the "test source," is described by density function $P(x)$ with

$$\text{probability} \quad [X \varepsilon A] = \int_A P(x)\, dx.$$

The other, called the "coded source," may be though of as the result of coding the output of a discrete sour with M symbols into the M code words. Thus

$$\text{probability} \quad [X = X^r] = \frac{1}{M} \quad \text{for} \quad r = 1 \cdots M.$$

All other vectors have probability zero.

The probability of error P_e for a given code is define by:

$$P_e = \frac{1}{M} \sum_{r=1}^M \int_{A_r} P(y/x^r)\, dy$$

where $A_r = \{y \mid P(y/x^s) \geq P(y/x^r)$ for some $s \neq r$

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[1] C. E. Shannon, "A mathematical theory of communication," *Bell Sys. Tech. J.*, vol. 27, pp. 379–423, 623–656; July/October, 1948.
[2] C. E. Shannon, "Certain results in coding theory for noisy channels," *Information and Control*, vol. 1, pp. 6–25; September, 1957.

For a given test source and channel, we can define ne mutual information function $I(x, y)$ by

$$I(x, y) = \frac{1}{N} \log_2 \frac{P(x, y)}{P(x)Q(x)}$$

here

$$P(x, y) = P(x)P(y/x), \qquad Q(x) = \int P(x, y)\, dx.$$

By plugging in the random variables, X and Y, we an get a random variable $I = I(X, Y)$. The expected alue of I, $\langle I \rangle$, is of course the rate of transmission of iformation, and the maximum value of $\langle I \rangle$ over-all test ources (for large N if the channel has memory) is the hannel capacity. As in Shannon's work,[2] the bound on ne error probability will be stated in terms of the distri-ution function of I,

$$\rho(\alpha) = \text{probability} \quad [I \le \alpha].$$

Additive Codes

If a and b are numbers in the interval (0, 1), define $+ b$ as their mod 1 sum; that is

$$a \oplus b = a + b \text{ if } a + b \le 1$$
$$= a + b - 1 \text{ if } a + b > 1.$$

imilarly, if u and v are vectors with components in the nterval (0, 1)

$$u \oplus v = (\mu_1 \oplus v_1, \cdots \mu_N \oplus v_N).$$

et $z^0, \cdots z^{NR}$ be vectors whose components are in the nterval (0, 1). Form 2^{NR} combination of the z's as follows

$$r = z^0 \oplus a_1 z^1 \oplus a_2 z^2 + \cdots a_{NR} z^{NR}, \qquad r = 1, \cdots 2^{NR}$$

here the a_s is the fifth binary digit in the binary expan-ion of the integer $r - 1$. Now we form our code words y applying a certain vector transformation (to be ex-lained below) to the w^r

$$x^r = g(w^r).$$

code formed in the above manner will be called an additive code." The Z's are called the "generators" f the code. Note that the whole procedure becomes ointless if the vector transformation $g(\cdot)$ is too compli-ated. For most channels this is not the case, however. n fact, we will see that for memory-free channels, the ransformation becomes a simple "stretching" function pplied to each component, *i.e.*,

$$x_s^r = h(w_s^r).$$

Random Additive Codes

If the generators are chosen at random, we have a random additive code;" mathematically we let the gene-ators be independent random vectors $Z^0, Z^1 \cdots Z^{NR}$ ach uniformly distributed in the N cube, $0 \le Z_s^t \le t$, $= 0 \cdots NR$, $s = 1 \cdots N$. This gives rise to random vectors W^r and X^r. It is easily verified that the W^r are uniformly distributed in the N cube, $0 \le W_s^r \le 1$, $r = 1 \cdots 2^{NR}$, $s = 1 \cdots N$, and are pairwise independent.

The function $g(\cdot)$ is used to match the code to the test source; *i.e.*, $g(\cdot)$ is chosen[3] so that for any set, A,

$$\text{probability} \quad \{X^r \in A\} = \int_A P(x)\, dx.$$

The X^r will of course also be pairwise independent.

For a random code the probability of error becomes a random variable and we can speak of the expected probability of error $\langle P_\epsilon \rangle$. We can also speak of the proba-bility that the random code words, X^r, meet certain conditions.

When using the coded source on a channel, (1) will be met if:

$$\frac{1}{M} \sum_{r=1}^{M} \frac{1}{N} \sum_{s=1}^{N} f(x_s^r) < f_{\max}. \tag{2}$$

If our code matches a test source which meets (1), the probability will be high when N is large that our random code words meet (2).

More precisely we have the following.

Theorem 1

If a random additive code matches an ergodic test source, $P(\mu)$, which meets (1), then the probability ap-proaches one with increasing N that the code words meet (2), provided $P(\mu)$ is such that $\langle f(X_s^r) \rangle$ and $\langle f^2(X_s^r) \rangle$ exist.

Proof

By the strong law of large numbers, $\langle f(X_s^r) \rangle < f_{\max}$. Let

$$\sigma^2 = \langle f^2(X_s^r) \rangle - \langle f(X_s^r) \rangle^2$$

and

$$S = \frac{1}{M} \sum_{r=1}^{M} \frac{1}{N} \sum_{s=1}^{N} f(X_s^r).$$

Now since the X_s^r are pairwise independent

$$\langle S^2 \rangle - \langle S \rangle^2 = \frac{1}{M^2 N^2} \sum_{r=1}^{M} \sum_{s=1}^{N} \sigma^2 = \frac{\sigma^2}{MN} \to 0$$

as N increases.

The theorem follows from Tchebycheff's inequality.

Theorem 2

If a random additive code matches a test source which yields a distribution of mutual information $\rho(\alpha)$, then

$$\langle P_\epsilon \rangle < \rho(R + \theta) + 2^{N\theta}$$

for any number θ.

[3] This is possible in a variety of ways. For example let $w_1 = F(x_1)$, $w_2 = F(x_2/x_1)$, $w_3 = F(x_3/x_1, x_2) \ldots$ when the F's are cumulative distributions. This obviously defines the x's as a function of the w's.

The proof follows essentially the method used by Shannon[2] and will be omitted here. The extension of Shannon's proof to the continuous case is trivial. The only other change to note is that if the probability of confusing the transmitted code word with a particular incorrect word is λ, then the total probability of error is $\leq 2^{NR}\lambda$. Shannon's random code words were independent and he used the inequality

$$(1 - \lambda)^M \geq 1 - M\lambda.$$

It is well known that, for a memory-free channel with constraint of the form (1), the channel capacity is obtained with a test source having independent symbols; *i.e.*,

$$P(x) = p(x_1)p(x_2) \cdots p(x_N).$$

The random variable I is then the average of independent random variable and by the weak law of large numbers

$$\rho(\alpha) \rightarrow 0 \quad \text{for} \quad \alpha < C.$$

Thus in Theorem 2, if $R < C$, set $\theta < C - R$ and $\langle P_\epsilon \rangle \rightarrow 0$.

Actually there is a much larger class of channels for which $\rho(\alpha) \rightarrow 0$ for $\alpha < C$ when $\rho(\alpha)$ is computed from the proper test source. This class includes channels with finite number of internal states along with many other more general ones, of course. Let us call such channels "regular." Then it is obvious that for any regular channel the expected probability of error can be made to approach zero if $R < C$.

Theorem 3

If a regular channel has a capacity C obtained with with an ergodic test source $P(\mu)$ and if $\langle f^2 \rangle$ and $\langle f \rangle$ exist [calculated with $P(\mu)$], then for any $\epsilon > 0$, $\delta > 0$, and any $R < C$, there exists an N and a random additive code of length N and signaling rate R such that the probability exceeds $1 - \epsilon$ that simultaneously the constraint (2) is met and the probability of error is less than δ.

Proof

The above remark show that the *expected* probability of error can be made arbitrarily small. Since the error probability is a positive quantity, this means that the probability can be made arbitrarily near one that the error probability is less δ. Choose N so large that the probability that $P_\epsilon < \delta$ is greater than $1 - \epsilon/2$ and also so large that the probability of meeting 2 is greater than $1 - \epsilon/2$; then the probability of failing either requirement is $< \epsilon$.

An Example

Let

$$P(y/x) = p(y_1/x_1) \cdots p(y_n/x_n)$$

where

$$p(y_1/x_1) = \frac{1}{\sqrt{2\pi}} e^{-(x_1 - y_1)^2/2}.$$

That is, the channel is memory-free with additive, whit Gaussian noise of unity power.

Let

$$f(x) = x^2, \qquad f_{\max} = 3.$$

The capacity of this channel is given by:

$$C = \tfrac{1}{2} \log (1 + 3) = 1 \quad \text{bit/sample.}$$

It is obtained by a test source having independent symbol with

$$P(x_1) = \frac{1}{\sqrt{6\pi}} e^{-x_1^2/6}.$$

Suppose we decide to signal over the channel at a rat of 0.75 bit per sample. Further assume that $N = 10$ turns out to be large enough to satisfy the various applica tions of the law of large numbers needed in the precedin theorem. Then with $R = 0.75$, $N = 100$ we need 7 generators of 100 components each. We therefore choos 7600 independent samples uniformly distributed in th interval (0, 1), z_s^r, $r = \theta \cdots 75$, $r = 1, \cdots 100$. Thes 7600 numbers are stored permanently at the transmitte and receiver. When the transmitter wishes to send block of 75 bits, it multiplies each of the z^r $r = 1, \cdots 7$ by the appropriate binary digit and adds the 76 vector mod one, to form a single vector of 100 component Each component, w_s $s = 1 \cdots 100$ is "stretched" b letting

$$x_s = h(w_s)$$

where $h(\cdot)$ is the inverse of the function $h^{-1}(\cdot)$,

$$h^{-1}(\mu) = \int_{-\infty}^{\mu} \frac{1}{\sqrt{6\pi}} e^{-x^2/6}\, dx.$$

This gives the μ_s a Gaussian distribution with the correc power. They are then transmitted to the receiver.

Unfortunately, there seems to be no simple detectio procedure. Apparently the receiver must compute eac of the 2^{75} possible transmitted combinations and compar them with the received signal. It is possible, howeve that the structure of these codes will be of some hel in designing a systematic detection procedure.

A Bibliography of Information Theory (Communication Theory—Cybernetics)*
(Third Supplement)

F. LOUIS H. M. STUMPERS†

Contents

)stracts:

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(No attempt has been made to list all available abstracts.)

** means that an English translation is available in the cor-sponding issue of the Pergamon Press (London) translated journal. *ektrosvyaz'* = *Telecommunications; Radiotekhnika* = *Radio ngineering*; and *Radiotekhnika i Elektronika* = *Radio Engineering d Electronics.*

‡ means that an English translation is available in the corre-onding issue of *Automation and Remote Control* (Consultant's ıreau, New York).

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VI. a) SPEECH. b) HEARING. c) VISION. d) LINGUISTICS, SEMANTICS. e) PERCEPTION (GENERAL) AND PERCEPTION (OTHER THAN VISUAL AND AURAL).

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VII. a) OTHER BIOLOGICAL APPLICATIONS (CYBERNETICS AND THE NERVOUS SYSTEM); b) HUMAN ENGINEERING; c) GROUP COMMUNICATION AND LEARNING.

a)

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A Relation Between Zero-Crossings and Fourier Coefficients for Band-width-Limited Functions*

In a recent paper,[1] an interpretation of a theorem by Titchmarsh[2] was discussed, showing that a function $f(z)$ whose Fourier transform was limited to the range $(-W, W)$ could be expressed as the infinite product

$$f(z) = f(0) \prod_{k=1}^{\infty} \left(1 - \frac{z}{z_k}\right) \qquad (1)$$

where

$$z_k = R_k e^{i\phi_k}$$

is the location of the kth zero (real or complex),

$$R_k \leq R_{k+1}$$

and

$$\lim_{k\to\infty} \frac{2WR_k}{k} = 1.$$

A special case of interest is

$$f(x) = x \prod_{k=1}^{\infty} \left(1 - \frac{n^2x^2}{\pi^2k^2}\right) \qquad (2)$$

which is the well-known product expansion for $\sin nx$. This function has uniformly-spaced zeros at $x = \pm k\pi/n$. By arranging the zeros in n repetitive groups displaced at intervals of π/n, the finite product expansion

$$\sin nx = 2^{n-1} \prod_{k=0}^{n-1} \sin\left(x + \frac{k\pi}{n}\right) \qquad (3)$$

is obtained heuristically.

The identity is easily vertified by showing that the finite product expansion:

1) is periodic in $2\pi/n$;
2) is an odd function of x; and
3) contains no frequency component higher than $n/2\pi$.

The constant 2^{n-1} is determined by expanding pairs of trigonometric products into sum and difference terms.

The result may be used to derive additional identities of interest as follows:

$$\log \sin nx = (n-1)\log 2 + \sum_{k=0}^{n-1} \log \sin\left(x + \frac{k\pi}{n}\right) \qquad (4)$$

$$\cot nx = \frac{1}{n} \sum_{k=0}^{n-1} \cot\left(x + \frac{k\pi}{n}\right) \qquad (5)$$

$$\csc^2 nx = \frac{1}{n^2} \sum_{k=0}^{n-1} \csc^2\left(x + \frac{k\pi}{n}\right) \qquad (6)$$

$$\prod_{k=0}^{n-1} \sin \frac{(1+2k)}{2n}\pi = \frac{1}{2^{n-1}} \qquad (7)$$

$$\prod_{k=1}^{n/2} \sin^2 \frac{k\pi}{n} = \frac{n}{2^{n-1}} \quad (n \text{ even}) \qquad (8)$$

$$\prod_{k=1}^{(n-1)/2} \sin^2 \frac{k\pi}{n} = \frac{n}{2^{n-1}} \quad (n \text{ odd}). \qquad (9)$$

The finite product expansion can be generalized to represent a periodic band-width limited function containing exactly n (real) zero crossings in the interval $(0, 2\pi)$. Thus, let

$$f(x_k) = 0; \qquad (0 \leq x_1 < \cdots < x_k < \cdots x_n \leq 2\pi). \qquad (10)$$

Then, except for an arbitrary scale factor, one may write

$$f(x) = \prod_{k=1}^{n} \cos \tfrac{1}{2}(x + \phi_k) \qquad (11)$$

* Received by the PGIT, September 28, 1959.

[1] F. E. Bond and C. R. Cahn, "On sampling the zeros of bandwidth-limited signals," IRE Trans. on Information Theory, vol. IT-4, pp. 110–113; September, 1958.

[2] E. C. Titchmarsh, "The zeros of certain integral functions," *Proc. London Math. Soc.*, vol. 25, pp. 283–302; May 14, 1926.

where

$$\phi_k = \pi - x_k$$

where the sine terms are replaced by cosine for subsequent mathematical convenience. A Fourier series representation for $f(x)$ in terms of its zero crossings may be obtained by expanding the finite product. Thus

$$f(x) = \frac{1}{2^{n-1}} \left\{ \cos\left(\frac{n}{2}x + \tfrac{1}{2}\Phi\right) + \sum_{k=1}^{n} \cos\left(\left[\frac{n}{2} - 1\right]x + \tfrac{1}{2}\Phi - \phi_k\right) + \sum_{\substack{k=2\\k>l}}^{n} \sum_{l-1}^{n} \cos\left(\left[\frac{n}{2} - 2\right]x + \tfrac{1}{2}\Phi - \phi_k - \phi_l\right) + \sum_{\substack{k=3\\k>l>m}}^{n} \sum_{l=2}^{n} \sum_{m=1}^{n} \cos\left(\left[\frac{n}{2} - 3\right]x + \tfrac{1}{2}\Phi - \phi_k - \phi_l - \phi_m\right) \cdots \right.$$

where

$$\Phi = \sum_{k=1}^{n} \phi_k. \qquad (12)$$

The amplitude and phase of each Fourier component is calculated by the summation of unit vectors indicated. For n odd, the function is repeated negatively in the range 2π to 4π, and the final term is the fundamental frequency component in the form $A \cos(x/2 + \theta)$. For n even, the final term is a constant and must be divided by two to correctly give the dc component. In either case, $n + 1$ phase and amplitude terms are present. However, the amplitude of the first term is independent of the locations of the n zero crossings.

The product expansion can be further generalized to include complex conjugate zeros in addition to real zero crossings. If the factors are considered in pairs

$$\cos \tfrac{1}{2}(x + \phi_k) \cos \tfrac{1}{2}(x + \phi_{k-1}) = \tfrac{1}{2}\left\{\cos\left(\frac{\phi_k - \phi_{k-1}}{2}\right) + \cos\left(x + \left[\frac{\phi_k + \phi_{k-1}}{2}\right]\right)\right\} = \tfrac{1}{2}\left\{\cos\left(\frac{x_k - x_{k-1}}{2}\right) - \cos\left(x - \left[\frac{x_k + x_{k-1}}{2}\right]\right)\right\}. \qquad (13)$$

Note that this factor vanishes at two values of x in $(0, 2\pi)$ and is specified in terms of the position of the mean of the zeros and half the distance between them. Now if x is generalized to the complex plane $z = x + jy$, and zeros occur at $z_k = x_k + jy_k$ and $z_{k-1} = x_k - jy_k$ then

$$\cos \tfrac{1}{2}(x + \phi_k) \cos \tfrac{1}{2}(x + \phi_{k-1}) = \tfrac{1}{2}(\cosh y_k - \cos[x - x_k]) \qquad (14)$$

in place of (13) for each pair of complex zeros.

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The Effect of a Limiter upon Signals in the Presence of Noise*

It has been shown[1] (the proof requires only a few lines) that when the input to a band-pass limiter is a narrow-band signal $A(t)\underline{/\theta(t)}$ plus narrow-band Gaussian noise of mean-square value σ^2 (both whithin the pass band of the device), the signal component of the otuput is

$$(kA/\sigma)_1F_1(\tfrac{1}{2}; 2; -A^2/2\sigma^2)\ \underline{/\theta(t)},$$

where k is a constant [$k = a/(2\pi)^{1/2}$ if the limiter output is a or 0 accordingly as its input is positive or negative] and $_1F_1$ is the confluent hypergeometric function. This result is obtained from (4) of Blachman[1] by setting $n = 0$ (the case of the ideal limiter) and $m = 1$ (the case of fundamental-frequency operation). Thus, the noisy band-pass limiter simply acts like a nonlinear device which changes amplitude A into $(kA/\sigma)\ _1F_1(\frac{1}{2}; 2; -A^2/2\sigma^2)$.

Now we may determine what nonlinear device of odd symmetry has this effect on amplitudes. Representing its characteristic by $f(x)$, we have

$$(kA/\sigma)_1F_1(\tfrac{1}{2}; 2; -A^2/2\sigma^2) = \int_0^{2\pi} f(A \cos\phi) \cos\phi\, d\phi/\pi.$$

Expanding both sides of this equation as power series in A and equating coefficients, we find that

$$f(x) = (kx/\sigma)_1F_1(\tfrac{1}{2}; 3/2; -x^2/2\sigma^2) = (2\pi)^{1/2}k(-\tfrac{1}{2} + \operatorname{erf} x/\sigma),$$

where $\operatorname{erf} x = \int_{-\infty}^{x} \exp -\frac{1}{2}t^2\, dt/(2\pi)^{1/2}$. Hence, the effect of the limiter upon signals within its pass band in the presence of noise is the same as that of an error-function-shaped nonlinearity in the absence of noise. It is for this reason that the action of an imperfect band-pass limiter with an error-function-shaped characteristic is especially easy to analyze;[2] its action is the same as that of an ideal band-pass limiter to whose input enough noise is added to obtain effectively the given error-function-shaped characteristic.

As a check on the foregoing results, we may rederive it *ab initio*. If the input to an ideal limiter is the sum of a signal $x(t)$ plus Gaussian noise $n(t)$ of mean-square value σ^2, both lying within its pass band, the output of the limiter before filtering being 0 or 1, according to the sign of the input, the instantaneous signal output voltage before filtering is simply the probability that $x + n$ is positive, *viz.*, erf x/σ. (The extra $-\frac{1}{2}$ above represents only a dc and would not be passed by the filter.) Hence, the signal output (including distortion) of the band-pass limiter is found by regarding the signal as passing through an error-function-shaped nonlinearity and then through a band-pass filter. To find the mean-square noise output, it is necessary to use the methods[3] of Blachman.[1]

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* Received by the PGIT, October 20, 1959.

[1] N. M. Blachman, "The output signal-to-noise ratio of a power-law device," *J. Appl. Phys.*, vol. 24, pp. 783–785; June, 1953. In (8), $I_0 - I_1$ should be changed in both places to $I_0 + I_1$.

[2] Janis Galejs, "Signal-to-noise ratios in smooth limiters," IRE Trans. on Information Theory, vol. IT-5, pp. 79–85; June, 1959.

[3] N. M. Blachman, "On Manasse, Price, and Lerner," IRE Trans. on Information Theory, vol. IT-4, p. 174; December, 1958.

On Best Approximation of Random Processes *et al.**

The purpose of this note is to point out that a theorem of E. Schmidt,[1] on the best L_2 approximation of a function of two variables by means of finite sums of products of functions of the first variable by functions of the second variable, can be extended rather simply to yield results on optimal signal representation[2] and coding for correlated signals.[3]

* Received by the PGIT, August 25, 1959.

[1] E. Schmidt, "Zur Theorie der linearen und nicht linearen Integralgleichungen," *Math. Ann.*, vol. 63, pp. 433–476; 1907.

[2] H. P. Kramer, "Best finite linear approximation to a second-order stochastic process," *Notices Amer. Math. Soc.*, vol. 5, p. 201; 1958.

[3] H. P. Kramer and M. V. Mathews, "A linear coding for transmitting a set of correlated signals," IRE Trans. on Information Theory, vol. IT-2, pp. 41–46; September, 1956.

For the sake of completeness, the expansion theorem and the approximation theorem shall be stated in the required generality. However, since their proofs, as presented for example in Smithies,[4] require no further elaboration than a reinterpretation in the more general context, they are omitted. The reinterpretation consists of allowing the first variable of a kernel to assume values in one space and the second variable to do so in a completely different space.

Let Ω_1, Ω_2 be two finite measure spaces with the respective measures μ_1 and μ_2. Let $L_2(\mu_1)$, $L_2(\mu_2)$ be the corresponding spaces of square-integrable functions. Let $K(s, t)$, with $s \varepsilon \Omega_1$, $t \varepsilon \Omega_2$, satisfy the following hypotheses.

a) $K(s, t)$ is measurable on $\Omega_1 \times \Omega_2$ and

$$\int_{\Omega_1 \times \Omega_2} | K(s, t) |^2 \, d(\mu_1 \times \mu_2)(s, t) < \infty. \tag{1}$$

b) For each value of s, $K(s, t)$ is a measurable function of t such that

$$\int_{\Omega_2} | K(s, t) |^2 \, d\mu_2(t) < \infty. \tag{2}$$

c) For each value of t, $K(s, t)$ is a measurable function of s such that

$$\int_{\Omega_1} | K(s, t) |^2 \, d\mu_1(s) < \infty. \tag{3}$$

d) $K(s, t)$ is not equivalent to the null function.

Under the above hypotheses it can be shown that there exists a singular system $(u_n, v_n; \lambda_n)$ of K defined by:

$$u_n(s) = \lambda_n \int_{\Omega_2} K(s, t) v_n(t) \, d\mu_2(s) \tag{4}$$

$$v_n(t) = \lambda_n \int_{\Omega_1} K(s, t) u_n(s) \, d\mu_1(s). \tag{5}$$

With respect to the singular system $(u_n, v_n; \lambda_n)$, in which u_n is normal in $L_2(\mu_1)$ and v_n is normal in $L_2(\mu_2)$ and $\lambda_n \geq \lambda_{n+1}$, the following L_2 expansion holds:

$$K(s, t) \doteq \sum_{n=1}^{\infty} \frac{u_n(s) v_n(t)}{\lambda_n} (L_2). \tag{6}$$

Moreover the following approximation theorem is valid.

Theorem (E. Schmidt[1]): Let $(u_n, v_n; \lambda_n)$ be a singular system for the L_2 kernel K and write

$$K_n(s, t) = \sum_{\nu=1}^{n} \frac{u_\nu(s) v_\nu(t)}{\lambda_\nu}. \tag{7}$$

If $a_1, \cdots, a_n, b_1, \cdots, b_n$ are arbitrary $L_2(\mu_1)$, $L_2(\mu_2)$ functions, then

$$\int_{\Omega_1 \times \Omega_2} \Big| K(s, t) - \sum_{\nu=1}^{n} a_\nu(s) b_\nu(t) \Big|^2 \cdot d(\mu_1 \times \mu_2)(s, t)$$
$$\geq \int_{\Omega_1 \times \Omega_2} | K(s, t) - K_n(s, t) |^2 \cdot d(\mu_1 \times \mu_2)$$
$$= \sum_{\nu=n+1}^{\infty} \frac{1}{\lambda^2}. \tag{8}$$

First Application

Let $x(\omega, t)$ be a second-order random function defined on the finite interval I of the real time axis. Upon identifying Ω_1 with Ω, the sample space of the process, and μ_1 with P the associated probability measure, and upon identifying Ω_2 with I and μ_2 with ordinary Lebesgue measure, one sees that (1) represents the Loève-Karhunen expansion of a second-order process and the approximation theorem implies the following.

Theorem:[2] Let $x(t)$ be a mean-square continuous stochastic process on I with zero mean. Let n be a positive integer. Let $\alpha_1, \alpha_2, \cdots, \alpha_n$ be n arbitrary random variables and $\varphi_1(t) \cdots, \varphi_n(t)$ n arbitrary functions. Then

$$\int_I E \left| x(t) - \sum_{\nu=1}^{n} \alpha_\nu \varphi_\nu(t) \right|^2 dt$$
$$\geq \int_I E \left| x(t) - \sum_{\nu=1}^{n} \frac{\xi_\nu \psi_\nu(t)}{\lambda_\nu} \right|^2 dt$$
$$= \sum_{\nu=n+1}^{\infty} \frac{1}{\lambda_\nu^2} \tag{9}$$

where ξ_ν are the random coefficients, and $\psi_\nu(t)$ the time functions appearing in the Loève-Karhunen expansion.[5] That is,

$$\text{a)} \quad \lambda_\nu^2 \int_I r(s, t) \overline{\psi_\nu(t)} \, dt = \psi_\nu(s), \tag{10}$$

$$\text{b)} \quad \xi_\nu = \lambda_\nu \int_I x(t) \overline{\psi_\nu(t)} \, dt, \tag{11}$$

$$\text{c)} \quad E\{\xi_i \bar{\xi}_j\} = \delta_{ij}, \quad \text{and} \tag{12}$$

$$\text{d)} \quad \int_I \psi_k(t) \overline{\psi l(t)} \, dt = \delta_k l. \tag{13}$$

Here $r(s, t)$ is the covariance function of $x(t)$.

It should be pointed out that this result is more general than a similar one presented by K. L. Jordan, Jr.,[6] in that the random variables $\alpha_1, \cdots, \alpha_n$ are arbitrary here whereas in Jordan[6] they are assumed to be the Fourier coefficients with respect to the system of functions $\varphi_1(t) \cdots, \varphi_n(t)$.

The above theorem suggests an optimal pulse amplitude modulation scheme. If it is desired for purposes of transmission to represent a time limited signal from a certain ensemble by means of a fixed finite set of numbers, then the best choice consists of the integrals of the signals against the singular functions $\{\psi_\nu(t)\}$. Having transmitted these numbers, an approximant to the original signal can then be reconstructed at the receiver by using the received numbers for coefficients in the truncated Loève-Karhunen expansion.

Second Application

Let Ω_1 be as in the first application, but identify Ω_2 with a finite or infinite set of integers and μ_2 with counting measure. Then one has the following theorem.

Theorem:[3] Let $x(\omega) = \{x_1(\omega), x_2(\omega), \cdots, x_m(\omega)\}$ be a random vector. If $a_1, a_2, \cdots, a_n$ is a sequence of random variables and $b^1, b^2, \cdots, b^n$ n arbitrary m-dimensional vectors, then

$$\sum_{r=1}^{m} E \left| x_r - \sum_{\nu=1}^{n} a_\nu b_r^\nu \right|^2$$
$$\geq \sum_{r=1}^{m} E \left| x_r - \sum_{\nu=1}^{n} \frac{\xi_\nu y_r}{\lambda_\nu} \right|^2$$
$$= \sum_{\nu=n+1}^{m} \frac{1}{\lambda_\nu^2}$$

where

$$\lambda_\nu^2 \sum_{s=1}^{m} R_{st} y_t^\nu = y_s^\nu$$

and R_{st} is the covariance matrix of x.

Conclusion

A random process can be regarded as a function of the two variables, time and position in phase space. On this premise, a theorem by E. Schmidt on the best least-squares "diagonal" representation of a function of two variables becomes applicable to random processes. The result is that, for a given random process $x(t)$, of all approximants having the form

$$\sum_{1}^{N} \alpha_n \varphi_n(t),$$

with N arbitrary random variables α_n and N arbitrary time functions $\varphi_n(t)$, the Loève-Karhunen representation is optimal. It is important to note the fact that it is not necessary to assume *a priori* that the α's are the Fourier coefficients corresponding to the φ's.

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[4] F. Smithies, "Integral Equations," Cambridge University Press, Cambridge, Eng., pp. 147–150; 1958.

[5] M. Loève, "Probability Theory," D. Van Nostrand Co., Inc., New York, N. Y., p. 464; 1955.

[6] K. L. Jordan, Jr., M.I.T. Res. Lab. Electronics, Cambridge, Mass., Quart. Progr. Rept. No. 52; January, 1959.

On the Cross Correlation Between Two Noisy Channels*

In this note we wish to derive a result for a two-channel configuration similar to one recently proved by Keilson, Mermin and Bello,[1] where mutually uncorrelated, Gaussian, narrow-band, channel inputs were assumed. Here, the bandwidth and correlation restrictions are removed and a more general nonlinear device is considered; specifically, the device is such that the output, $v(t)$, is given in terms of the input, $u(t)$, by the relationship:

$$v(t) = \int_0^\infty K[u(t-\tau), \tau]\, d\tau, \qquad (1)$$

where $K(x, y)$ is a real-valued function of its arguments satisfying the one-sided homogeneity condition,

$$K(\lambda x, y) = \lambda^\nu K(x, y) \qquad (2)$$

for all $\lambda > 0$ and some real ν This class of devices is a subclass of Zadeh's[2] class $\mathfrak{N}_1$ and contains such well-known (instantaneous) devices as the ideal limiter ($\nu = 0$), full and half-wave nth law devices ($\nu = n$) and any tandem conbination of the foregoing with a linear filter.

Consider two channels, each containing a nonlinear device characterized by (1) and (2), the devices being identical for the two channels. Let the input to Channel 1 be $N_0(t) + N_1(t)$ and the input to Channel 2 be $N_0(t) + N_2(t)$, where the $\{N_i(t)\}$, ($i = 0, 1, 2$) are mutually correlated, stationary, zero-mean Gaussian noises; the problem then is the determination of the output cross correlation[3]

$$\psi_{12}(\tau) = \langle v(t)v(t+\tau)\rangle.$$

From (1) and the assumed input forms,

$$\psi_{12}(\tau) = \int_0^\infty \int_0^\infty \langle K[N_0(t-\xi) + N_1(t-\xi), \xi]\cdot K[N_0(t+\tau-\eta) + N_2(t+\tau-\eta), \eta]\rangle\, d\xi\, d\eta. \qquad (3)$$

Letting

$$x_1 = N_0(t-\xi) + N_1(t-\xi)$$

and

$$x_2 = N_0(t+\tau-\eta) + N_1(t+\tau-\eta),$$

the joint probability density of x_1 and x_2 is[4] $P(x_1, x_2; \alpha, g(\tau-\eta+\xi), \beta)$, where

$$P(x_1, x_2; a, b, c) \triangleq \frac{1}{2\pi\sqrt{ac-b^2}} \cdot \exp\left[-\frac{cx_1^2 - 2bx_1x_2 + ax_2^2}{2(ac-b^2)}\right], \qquad (4)$$

$$\alpha = \langle x_1^2(t)\rangle, \qquad \beta = \langle x_2^2(t)\rangle$$

and

$$g(\tau) = \langle x_1(t)x_2(t+\tau)\rangle.$$

From (4), it is apparent that $P(x_1, x_2; \sigma^2, \sigma^2\rho(\tau), \sigma^2) \triangleq W(x_1, x_2; \sigma^2, \rho(\tau))$ is the usual second-order probability density for a zero-mean, stationary Gaussian process with rms σ and normalized autocorrelation function $\rho(\tau)$. Further, the identity

$$P(x\sqrt{\alpha}, y\sqrt{\beta}; \alpha, \alpha\beta h(\tau), \beta) \equiv \frac{W(x, y; 1, h(\tau))}{\sqrt{\alpha\beta}} \qquad (5)$$

is easily verified. Eq. (3) may be rewritten

$$\psi_{12}(\tau) = \int_0^\infty \int_0^\infty \left[\int_{-\infty}^\infty \int_{-\infty}^\infty K(x_1, \xi) \cdot K(x_2, \eta) P(x_1, x_2; \alpha, \cdot g(\tau-\eta+\xi), \beta)\, dx_1\, dx_2\right] d\xi\, d\eta. \qquad (6)$$

Under the change of variable $x = x_1/\sqrt{\alpha}$, $y = x_2/\sqrt{\beta}$ and the definition

$$h(\tau) = \frac{g(\tau)}{\sqrt{\alpha\beta}},$$

(6) becomes

$$\psi_{12}(\tau) = (\alpha\beta)^{\nu/2} \cdot \int_0^\infty \int_0^\infty \left[\int_{-\infty}^\infty \int_{-\infty}^\infty K(x,\xi)K(y,\eta) \cdot W(x, y; 1, \cdot h(\tau-\eta+\xi))\, dx\, dy\right] d\xi\, d\eta, \qquad (7)$$

where use has been made of (5) and relation (2). If we define $\phi_\rho(\sigma, \tau)$ to be the output autocorrelation function of a single nonlinear device with characteristic $K(x, y)$ for a stationary, zero-mean, Gaussian input having rms σ and normalized autocorrelation function $\rho(\tau)$, then the quadruple integral in (16) is equivalent to the quantity $\phi_h(1, \tau)$. Thus

$$\psi_{12}(\tau) = (\alpha\beta)^{\nu/2}\phi_h(1, \tau), \qquad (8)$$

which is the main result of this note.

For the subclass of zero-memory devices satisfying (1) and (2) and for mutually uncorrelated $N_i(t)$ ($i = 0, 1, 2$), (8) can be shown to reduce to

$$\psi_{12}(\tau) = \frac{1}{Z}\phi_{Z\rho_{00}}(\sigma_0^2) \qquad (9)$$

where σ_0 and ρ_{00} are respectively the rms and normalized aurocorrelation function for the process $N_0(t)$, and

$$Z = \left\{\left(1 + \frac{\sigma_1^2}{\sigma_0^2}\right)\left(1 + \frac{\sigma_2^2}{\sigma_0^2}\right)\right\}^{-1/2}.$$

Eq. (9) is a result which is formally analogous to (12) of the previously mentioned work.[1]

Two of the explicit examples cited previously[1] are included in the formulation of (8). The square-law detector may be obtained from the tandem combination of an instantaneous square device and low-pass (linear) filter while the envelope detector is obtained by a half-wave linear device followed by a low-pass filter.

From the proof of (8), it is apparent that the result may be generalized to include the case of P additive zero-mean, stationary, Gaussian noises in Channel 1 and Q such inputs in Channel 2 with arbitrarily given correlations existing among the various noises. The result will again have the form of (8); however, obvious modifications must be made in the definitions of α, β, and $h(\tau)$.

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* Received by the PGIT, September 21, 1959.

[1] J. Keilson, N. D. Mermin, and P. Bello, "A theorem on cross correlation between noisy channels," IRE Trans. on Information Theory, vol. IT-5, pp. 77–79; June, 1959.

[2] L. A. Zadeh, "Optimum nonlinear filters," *J. Appl. Phys.*, vol. 24, pp. 396–404; April, 1953.

[3] Angle brackets denote ensemble averaging.

[4] J. H. Laning and R. H. Battin, "Random Processes in Automatic Control," McGraw-Hill Book Co., Inc., New York, N. Y., p. 78 and p. 156; 1956.

Probability Distribution for the Derivative of the Envelope of Signal and Gaussian Noise*

In a recent paper, Brown[1] demonstrated that the first probability distribution for the derivative of the envelope of Gaussian noise is Gaussian. This note serves to point out further conditions under which this property prevails. If a constant amplitude, constant frequency "carrier" signal is linearly added with the input Gaussian noise prior to the detector action, then the first probability distribution for the derivative of the envelope of such a process will be Gaussian, provided that the frequency spectrum of the input noise possesses narrow-band symmetry about the carrier frequency. This may be readily shown in the expression for the characteristic function whose development may be obtained in a simple, straightforward fashion as briefly reviewed in the following.

If we express the signal applied to the input of the envelope device as

$$e(t) = S_0 \cos\omega_0 t + x(t)\cos\omega_0 t + y(t)\sin\omega_0 t \qquad (1)$$

* Received by the PGIT, August 31, 1959.

[1] W. M. Brown, "Some results on noise through circuits," IRE Trans. on Information Theory, vol. IT-5, pp. 217–227; May, 1959.

where S_0 = amplitude of carrier signal, $x(t)$ = inphase noise component, and $y(t)$ = quadrature noise component; then the envelope function becomes

$$z(t) = \sqrt{[x(t) + S_0]^2 + y^2(t)}. \tag{2}$$

The derivative of z becomes

$$\dot z(t) = \frac{(x + S_0)\dot x + y\dot y}{\sqrt{(x + S_0)^2 + y^2}}. \tag{3}$$

The characteristic function associated with the first probability density function of $\dot z(t)$ is given by

$$\Psi_{\dot z}(\Omega) = \iiiint_{-\infty}^{\infty} e^{-i\Omega\dot z} \cdot f(\dot x, \dot y, x_1, y)\, d\dot x\, d\dot y\, dx_1\, dy \tag{4}$$

where $x_1 = x + S_0$.

If the input Gaussian noise is a stationary process, then it may be shown that

$$f(\dot x, \dot y, x_1, y) = f(\dot x, y) f(\dot y, x_1) \tag{5}$$

where each f is a Gaussian probability density function. The following covariance matrices are associated therewith:

$$\begin{Vmatrix} \sigma_2^2 & \nu\sigma_1\sigma_2 \\ \nu\sigma_1\sigma_2 & \sigma_1^2 \end{Vmatrix} \quad \text{and} \quad \begin{Vmatrix} \sigma_2^2 & -\nu\sigma_1\sigma_2 \\ -\nu\sigma_1\sigma_2 & \sigma_1^2 \end{Vmatrix}$$

where

$$\sigma_1^2 = \frac{1}{2\pi}\int_{-\omega_0}^{\infty} W(\omega')\, d\omega' = \overline{(x_1 - S_0)^2} = \overline{y^2} \tag{6a}$$

$$\sigma_2^2 = \frac{1}{2\pi}\int_{-\omega_0}^{\infty} \omega'^2 W(\omega')\, d\omega' = \overline{\dot x^2} = \overline{\dot y^2} \tag{6b}$$

$$\nu\sigma_1\sigma_2 = \frac{1}{2\pi}\int_{-\omega_0}^{\infty} \omega' W(\omega')\, d\omega' = \overline{\dot x y} = -\overline{\dot y(x_1 - S_0)}. \tag{6c}$$

$W(\omega)$ equals the spectral density of the input noise and $\omega' = \omega - \omega_0$. The joint distribution function $f(\dot x, \dot y, x_1, y)$ accordingly becomes

$$f(\dot x, \dot y, x_1, y) = \frac{1}{(2\pi)^2\sigma_1^2\sigma_2^2(1 - \nu^2)} \cdot \exp - \frac{1}{2(1 - \nu^2)}\left[\frac{\dot x^2}{\sigma_2^2} + \frac{\dot y^2}{\sigma_2^2} + \frac{(x_1 - S_0)^2}{\sigma_1^2} + \frac{y^2}{\sigma_1^2} - \frac{2\nu\dot x y}{\sigma_1\sigma_2} + \frac{2\nu\dot y(x_1 - S_0)}{\sigma_1\sigma_2}\right]. \tag{7}$$

After substituting (7) into (4), we can integrate with respect to $\dot x$ and $\dot y$ to obtain

$$\Psi_{\dot z}(\Omega) = \left\{\exp\left[-\frac{\Omega^2}{2}\sigma_2^2(1 - \nu^2)\right]\right\} \cdot \left\{\int_0^{\infty}\int_0^{2\pi} \frac{r}{2\pi\sigma_1^2}\exp\left[-\frac{r^2 + S_0^2}{2\sigma_1^2}\right] \cdot \exp\left[\frac{S_0}{\sigma_1^2}(r\cos\theta - i\nu\sigma_1\sigma_2\Omega\sin\theta)\right] d\theta\, dr\right\} \tag{8}$$

where we have changed coordinates in accordance with

$$x_1 = r\cos\theta \quad \text{and} \quad y = r\sin\theta.$$

Integrating with respect to θ yields

$$\Psi_{\dot z}(\Omega) = \exp\left[-\frac{\Omega^2}{2}\sigma_2^2(1 - \nu^2)\right] \cdot \int_0^{\infty} \mu \exp\left[-\frac{\mu^2 + \alpha^2}{2}\right] \cdot I_0(\alpha\sqrt{\mu^2 - k^2})\, d\mu \tag{9}$$

where

$$\frac{r}{\sigma_1} = \mu; \qquad \frac{S_0}{\sigma_1} = \alpha \quad \text{and} \quad k = \nu\Omega\sigma_2.$$

Brown's conclusions are demonstrated by letting $S_0 = 0$ (and hence $\alpha = 0$), to obtain for $\Psi_{\dot z}(\omega)$ the following:

$$\Psi_{\dot z}(\Omega) = \exp\left[-\frac{\Omega^2\sigma_2^2}{2}(1 - \nu^2)\right] \tag{10}$$

which is, as is well known, the characteristic function for a normal distribution of zero mean and variance $\sigma_2^2(1 - \nu^2)$.

When the frequency spectrum of the input noise has symmetry about the carrier frequency, then by (6c), $\nu = 0$ (and hence $k = 0$). Accordingly, the integral expression (9) will not be a function of Ω; it will, in fact, be unity for all "α." Therefore, for such a case

$$\Psi_{\dot z}(\Omega) = \exp\left[-\frac{\Omega^2}{2}\sigma_2^2\right]. \tag{11}$$

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Contributors

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Bernard Elspas (S' 46—A' 48—S' 53—A' 55—M' 56) was born in New York, N. Y., on July 26, 1925. He received the B.E.E. degree from the College of the City of New York in 1946, the M.E.E. degree from New York University, New York, in 1948, and the Ph.D. degree from Stanford University, Stanford, Calif., in 1955.

From 1949 to 1951 he was a research assistant and a research associate with New York University's Electron Tube Group. While studying for the Ph.D. degree at Stanford, he was a part-time research assistant at the Electronics Research Laboratory of the University. Upon completing his degree requirements, Dr. Elspas remained at Stanford in the Applied Electronics Laboratory studying the application of statistical communication theory research in radar systems. He also taught courses in communication theory for the University of California Extension Program. In 1955 he joined the Stanford Research Institute, Menlo Park, Calif., where he was engaged in the development and testing of the ERMA (Electronic Recording Machine, Accounting) computer. Since then he has conducted research in network synthesis and the organization of communication systems.

He is currently engaged in a theoretical investigation of the analysis and synthesis of sequential digital systems.

Dr. Elspas is a member of Eta Kappa Nu, Sigma Xi, and RESA. He has been the recipient of the Belden Mathematical Award, the New York State Regents Scholarship, and the National Science Foundation Fellowship.

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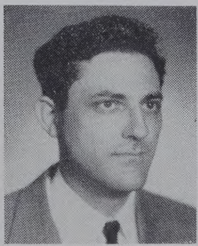

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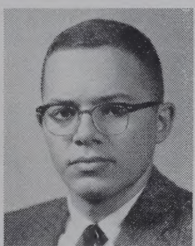
❖

Joel Max was born in New York, N. Y., on April 27, 1934. He received the degree of Bachelor of Engineering Physics from Cornell University, Ithaca, N. Y., in 1957, and the M.S. degree in electrical engineering from Massachusetts Institute of Technology, Cambridge, in 1959.



J. MAX

While at M.I.T. he was associated with the Research Laboratory of Electronics in Cambridge, and the Lincoln Laboratory in Lexington, Mass. He continued his association with the Lincoln Laboratory after he received his degree, and was principally engaged in research on information and communication theories and communication systems. He is currently doing research at the Imperial College of Science and Technology, London, England, as a Fulbright Fellow.

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He came to the United States in 1947 as Technical Consultant for the United States Air Force. He was with the Air Force Cambridge Research Center, Cambridge, Mass., until January, 1959 and was engaged there in work on antennas, propagation, and communication techniques.

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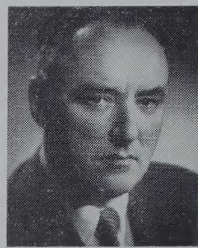

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In 1939 he joined the Department of Fundamental Radio Research under Professor van der Pol. In this group he did work on circuit theory, cables, frequency modulation, stochastic problems, and noise. In 1946 he received the doctoral degree in technical science from Delft University. Since then he has done work on low-noise receivers, general noise problems and information theory, and radio interference. In 1952–1953, on leave of absence from the Philips Laboratories, he was a research associate at the Research Laboratory of Electronics of the Massachusetts Institute of Technology, Cambridge.

From its conception in 1948, Dr. Stumpers has been a member of the board of the Foundation for the Study of Radioastronomy in The Netherlands. He is a member of the Dutch Physical Society, the Dutch Radio Society, and a recipient of the Veder Radio Prize. He is also chairman of the committee for information theory of URSI, vice president of the Netherlands CISPR Committee, and a member of the CISPR Steering Committee. He also attended the CCIR Conferences in Warsaw and Los Angeles as an expert for the Netherlands delegation.

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